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ELEKTOR ELECTRONICS

21 PHONE NUMBERS

THE INTERNATIONAL ELECTRONICS MAGAZINE

July/August 1994
£ 3.50

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ELEKTOR ELECTRONICS

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- PIC programming course - Part 2
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COMPUTERS & MICROPROCESSORS

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POWER SUPPLIES & BATTERY CHARGERS

RADIO, TELEVISON & COMMUNICATIONS

TEST & MEASUREMENT

MISCELLANEOUS INFORMATION

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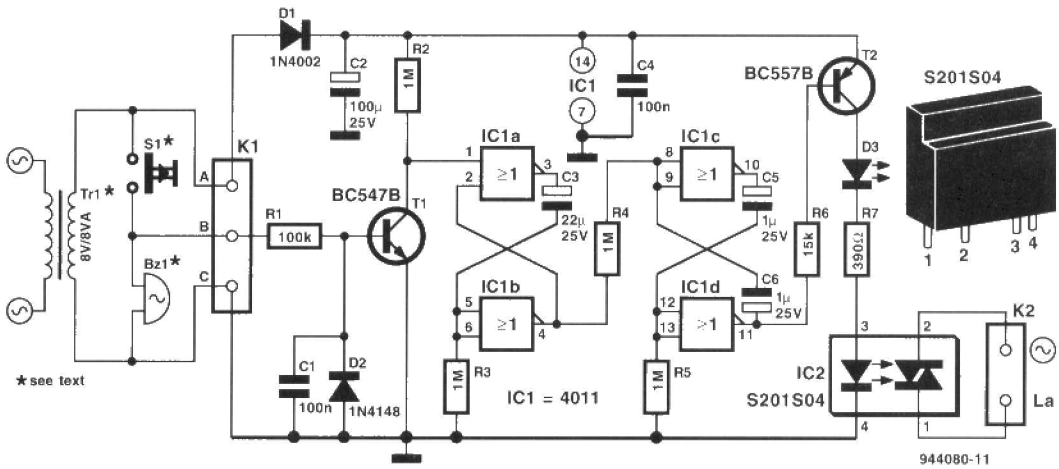
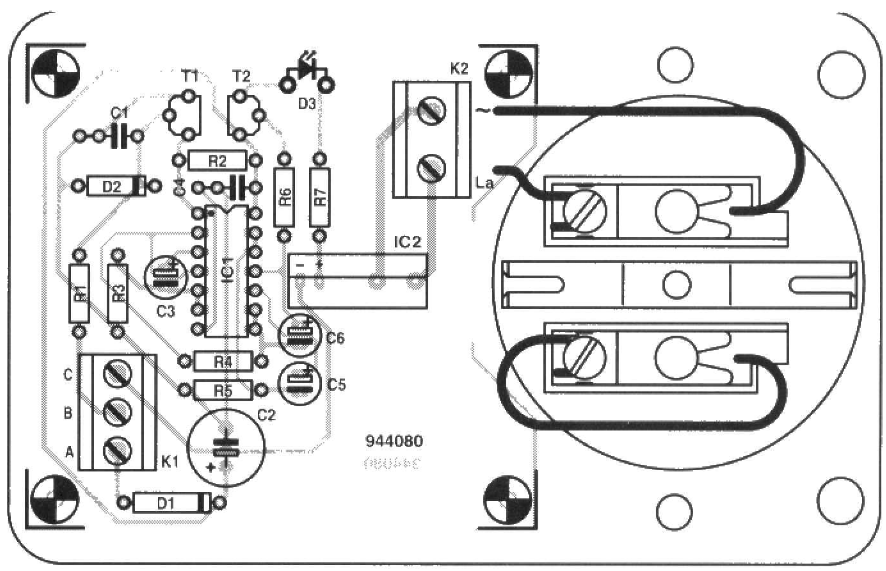
OPTICAL DOORBELL

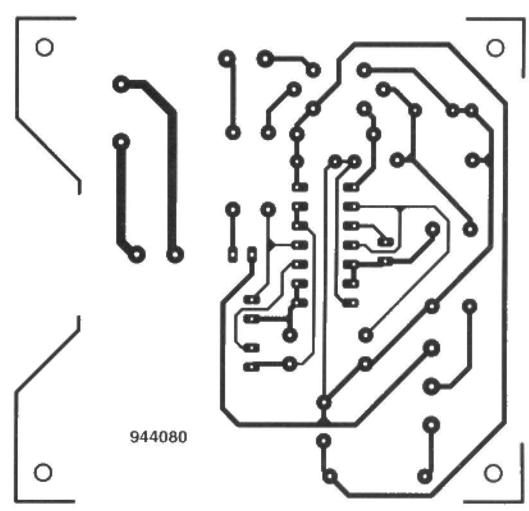
This little circuit is for all cases where you can not hear the doorbell, for whatever reason, but would still like to be alerted if someone calls at you. The reason for being unable to notice the sound of the bell may be that you are hard of hearing, blown away by an old Jesus & Mary Chain song, or deafened by the noise of the circular saw in your workshop as it grinds its way through a piece of wood or metal. Whatever the case, someone wants to see you, and an optical warning is a godsend.

The circuit uses the existing bellpush (S_1) and associated transformer (Tr_1), which is usually rated at 8 V, 1 A, and safe to use for the present application. The secondary voltage is rectified and smoothed by D_1-C_2 which power the optical doorbell driver. When the push-button is pressed, the bell will ring as usual. At the same time, transistor T_1 is switched on and off at the rate of the mains frequency (50 or 60 Hz). This, in turn, causes a monostable multivibrator formed by IC_{1a} and IC_{1b} to be started. The monotime is set to a value of about 15 seconds by C_3 and R_3. The monostable in turn enables an oscillator, IC_{1c}-IC_{1d}, which controls output driver transistor T_2. Consequently, the lamp connected to the solid-state relay, IC_2, will flash for a predetermined period. LED D_3 flashes at the same rate as a means of checking the operation of the optical door bell. The maximum power of the bulb connected to the circuit is about 100 W.

Since the mains voltage is present on two copper tracks and four solder points on the printed circuit board, the circuit must be built with due attention paid to electrical safety. The completed printed circuit board is built into a solid ABS enclosure with integral mains socket, so that the bulb can be connected via an ordinary mains cable and plug.







Parts list

Resistors:

R_1 = 100 kΩ
R_2–R_5 = 1 MΩ
R_6 = 15 kΩ
R_7 = 390 Ω

Capacitors:

C_1,C_4 = 100 nF
C_2 = 100 µF, 25 V, radial
C_3 = 22 µF, 25 V, radial
C_5,C_6 = 1 µF, 25 V, radial

Semiconductors:

D_1 = 1N4002
D_2 = 1N4148
D_3 = LED
T_1 = BC547B
T_2 = BC557B

Integrated circuits:

IC_1 = 4011
IC_2 = S201S04 (Sharp)

Miscellaneous:

K_1 = 3-way PCB terminal block, pitch 5mm.
K_2 = 2-way PCB terminal block, pitch 7.5mm.
S_1 = existing bellpush.
Tr_1 = existing bell transformer 8V/1A.
BZ_1 = existing doorbell.
Enclosure: e.g., Bopla Nicro N12. [Phoenix Mecano UK Ltd., Tel. (0296) 398 853.]
PCB Ref. 944080 (p. 110).

Design: E. Verbeek
[944080]

RC5 TRANSMITTER WITH 80C535

The 80C535 SBC (single board computer) can be adapted to function as an RC-5 infra-red (IR) transmitter by the addition of some hardware and software described in this article.

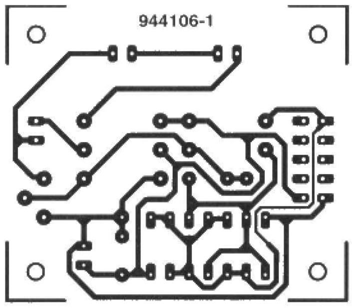
The required carrier wave of 36 kHz is generated by software: it is, therefore, essential that a 12 MHz crystal is used as the clock for the microcontroller.

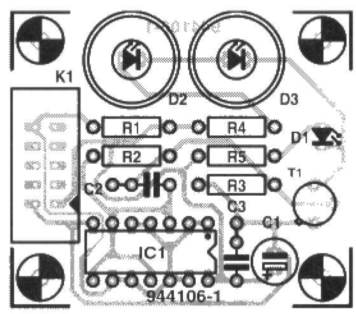
The circuit consists of four parallel-connected buffers Type 74HCT00, followed by a Type 2N222 transistor, T_1, which drives IR transmit diodes D_2 and D_3. The edges of the transmitted signal are enhanced by C_2.

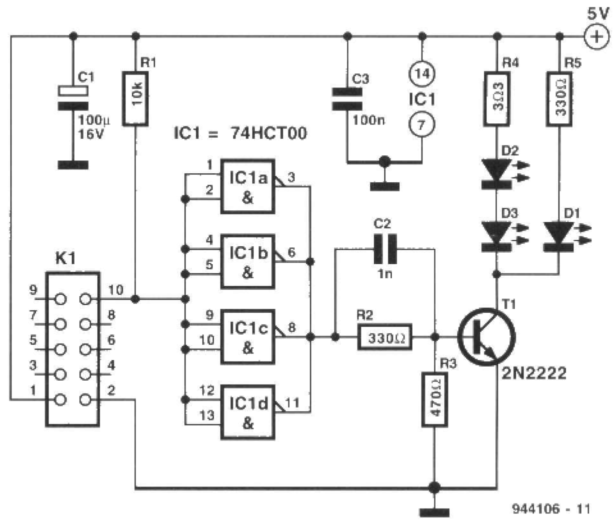
The IR transmit diodes convert the digital code into infrared signals. Light-emitting diode D_1 gives a visibile indication that a code is being transmitted.

The connection with the SBC is made via K_1.

The circuit is best built on the printed-circuit board that, together with the necessary software, is available through our Readers' Services.







Parts list

Resistors:
R_1 = 10 kΩ
R_2, R_5 = 330 Ω
R_3 = 470 Ω
R_4 = 3.3 Ω

Capacitors:
C_1 = 100 µF, 10 V, radial
C_2 = 1 nF
C_3 = 100 nF

Semiconductors:
D_1 = LED, 3 mm, red
D_2, D_3 = LD271
T_1 = 2N222

Integrated Circuits:
IC_1 = 74HCT00

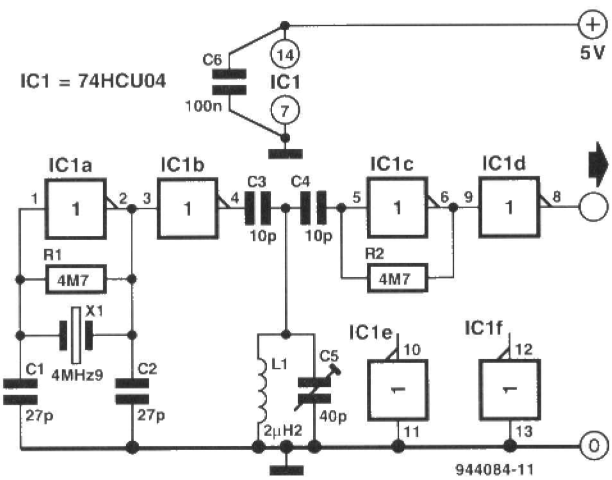
Miscellaneous:
K_1 = 10-way straight box header
PCB Ref. 944106 (p. 110)
Software Ref. 946199 (p. 110)

Design: W. Hackländer and S. Furchtbar
[944106]

OVERTONE OSCILLATOR

Quartz crystals are ground by the manufacturers to oscillate either on the fundamental frequency or on one of the odd harmonics (overtones). Now-adays, this grinding is so accurate that fundamental mode crystals do not oscillate on the third or fifth overtone (as they usually could do in years past). However, the present circuit enables fundamental mode crystals to oscillate on an overtone (third or fifth).

In the design, use is made of the fact that rectangular waveforms contain odd harmonics. The signal generated by oscillator IC_{1a} is amplified in IC_{1b}. This means that the edges of the signal become steeper, which gives the signal more harmonics. The wanted overtone is selected by resonant circuit L_1-C_5, amplified by IC_{1c} and shaped into a proper square wave by inverter IC_{1d}.



In the prototype, a fundamental mode crystal of 4.9 MHz was used and this oscillated unfailingly on the third as well as on the fifth overtone. The circuit can work on other frequencies, but the value of L_1 may then have to be altered by trial and error.

The circuit draws a current of only a few milliamperes.

Design: K. M. Walraven
[944084]

SURROUND SOUND

The quality of reproduced sound does not result from a single property of the audio signal, but from the sum total of several characteristics. It is not only the frequency response, the signal-to-noise ratio and distortion, but also the breadth and depth of the stereo image that determine the degree to which the reproduced sound is experienced as faithful. The breadth of the image is, within certain limits, under the listener's control (by placing the loudspeakers in the desired positions). However, since the dimensions of an average living room and a concert hall are vastly different, obtaining the right depth of the image is rather more difficult. There are now amplifiers on the market that provide 'surround sound', that is, sound that appears to surround the listener as it does in a concert hall. It is fortunately not necessary to buy one of these modern amplifiers, since the effect can also be obtained with an existing amplifier as this article shows.

Design by T. Giesberts

Design considerations

The ultimate way of producing surround sound is to make a four-channel recording and play this back via four separate channels: two for speakers in front of the listener and two for speakers behind him. Several manufacturers introduced this quad system some years ago, but it was not a commercial success. This was mainly because of the high cost: the system required a complex decoder, a second stereo amplifier and two extra speakers.

Then, there was a lot of experimentation by manufacturers with matrix circuits that produced a sort of quasi-quad sound from a normal stereo signal. The results were encouraging, but the associated costs were almost as high as those of the real quad systems.

Furthermore, there is the possibility of driving two 'rear speakers' by an 'after sound' apparatus. This produces interesting sounds, but requires quite a lot of electronic circuitry.

Many people create surround sound by connecting two extra loudspeakers, placed at the back of the listener, in parallel with the existing front loudspeakers. However, this does not really create surround sound, since the rear loudspeakers sound exactly the same as the front ones.

Finally, the so-called L-R signal may be used to drive the rear loudspeaker(s). This yields very satisfactory results and is relatively inexpensive to achieve. This technique is used in the circuit described. Although the technique is not new, its present application, to the best of our knowledge, is.

L-R signal

The L-R (or R-L) signal is exactly what its name indicates: the difference between the left-hand and right-hand channels. Note, however, that only the signals that are exclusive to the left-hand or the right-hand channel are used. Signals that occur equally in the left-hand and the right-hand channel are not represented in the difference signal.

As is well-known, a stereo signal below about 200 Hz has hardly any directivity. This does not matter, fortunately, because those low frequencies are spread more or less uniformly in all directions by the loudspeakers. In other words, at such low frequencies, the reproduced sound is already 'surround sound'.

The situation is quite different at mid- and high frequencies, which are far more directional. As it so happens, it is just these frequencies above 200 Hz that are contained either in the left-hand or in the

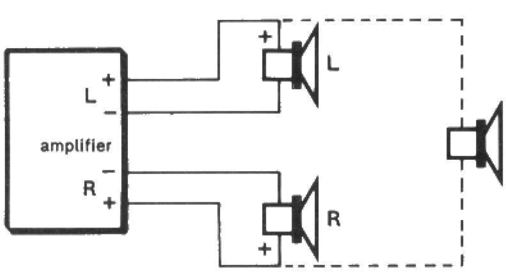


Fig. 1. An additional loudspeaker connected as shown will reproduce only the difference between the two stereo channels. The difference is called the L-R (or R-L) signal.

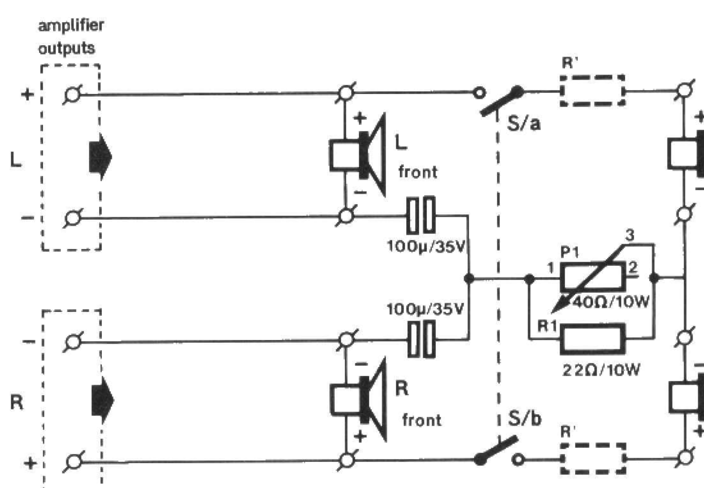


Fig. 2. The final design. The effect of the loudspeakers at the rear of the listener can be adjusted with P_1 from 'L-R' to 'double stereo'. If the rear speakers are too loud, they can be moderated by resistors *R*.

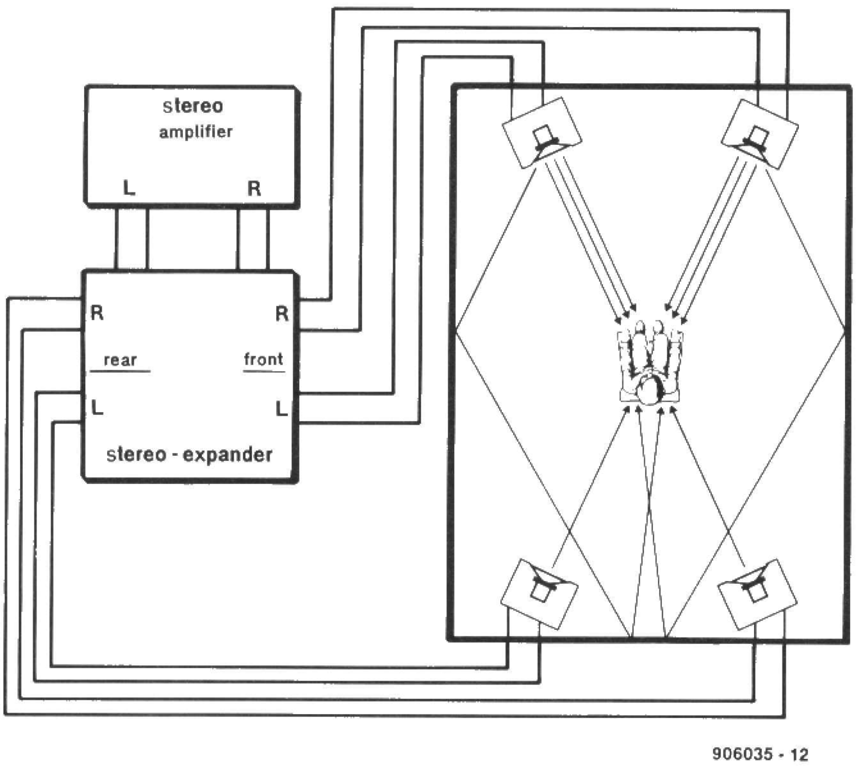


Fig. 3. Adding a small circuit and two (inexpensive) loudspeakers to an existing audio system gives a surround sound system.

right-hand channel, **not** in both.

It is clear from this that the difference between the two channels contains just the right information to drive one or two loudspeakers at the rear of the listener.

Figure 1 shows that the L-R signal is obtained by connecting the extra loudspeaker between the two + loudspeaker terminals of the amplifier. The signals below 200 Hz (which are identical in both channels) thus appear both in phase and in anti-phase across the third speaker and so cancel one another.

The setup in **Fig. 1** could be used in a practical application, were it not that it is rather simplistic.

Final configuration

Figure 2 shows the final setup, which is basically the same as that in **Fig. 1**, but there are now four loudspeakers. It will be seen that the additional (rear) speakers are connected in series across the two + loudspeaker terminals of the amplifier. The extra loudspeakers can be switched on and off with S_1.

Since sound affects different people differently, and it is, therefore, not certain whether the pure L-R effect will please everyone, adaptation to individual taste is made possible by P_1. If this control is set to maximum resistance, the effect is almost the same as if P_1 were not there (that is, the loudspeakers reproduce only the L-R signal). With P_1 set to about the centre of its travel, a sort of spatial stereo is produced, while with the control at minimum resistance the rear and front speakers are in parallel, which results in a sort of 'double stereo'. In the latter case, C_1 and C_2 ensure that the (superfluous) low frequencies (below 200 Hz) are not reproduced by the rear speakers. These bipolar electrolytic capacitors also prevent any earth loops.

The potentiometer has a value of about 40 Ω and a rating of 10 W. Shunt resistor R_1 increases the load capacity to some extent and gives the control a somewhat refined character. If a suitable potentiometer can not be obtained, a so-called L-pad can be used (as was done in the prototype).

Suitable drive units

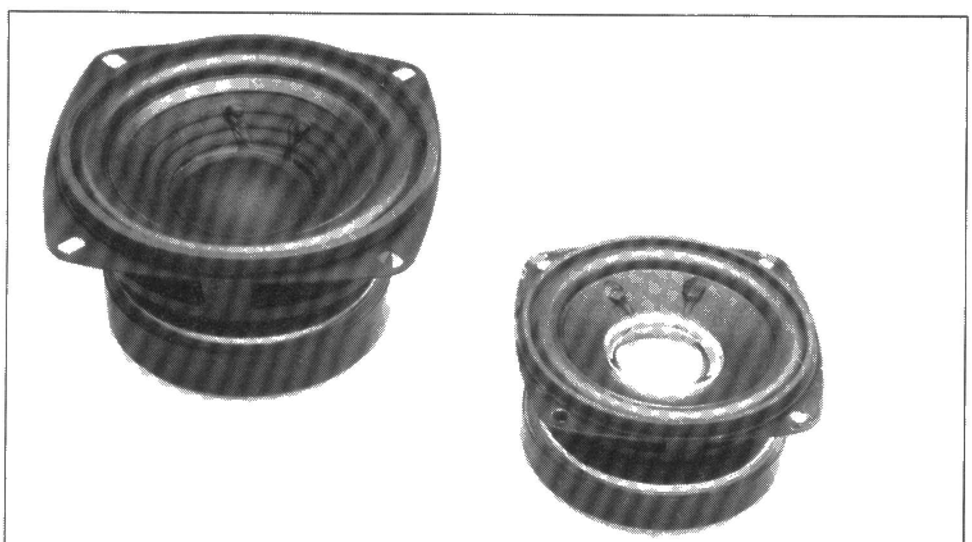
As has already been mentioned, the rear loudspeakers do not have to reproduce low frequencies below about 200 Hz. This means that large enclosures are not necessary, because these are needed only for bass reproduction. In practice, it is found that frequencies above 5000 Hz also do not need to be reproduced by the rear loudspeakers. All this comes down to the fact that the rear speakers can be made from a good medium-frequency or wide-band drive unit in an enclosure with a volume of not more than 2–3 litres. Car radio speakers of the better variety are eminently suitable.

It is important, however, that the efficiency of the rear speakers is not too high, since, to prevent the effect becoming too emphatic and unnatural, these speakers should be only just audible. This is why in **Fig. 2** resistors *R* are shown in dashed lines: they are for use if the rear speakers need to be moderated to some extent. Their value must be determined empirically, but will normally be 2.2–10 Ω (10 W).

Finally

It is advisable to fit the various components in a small case, which is connected between amplifier and loudspeakers as shown in **Fig. 3**.

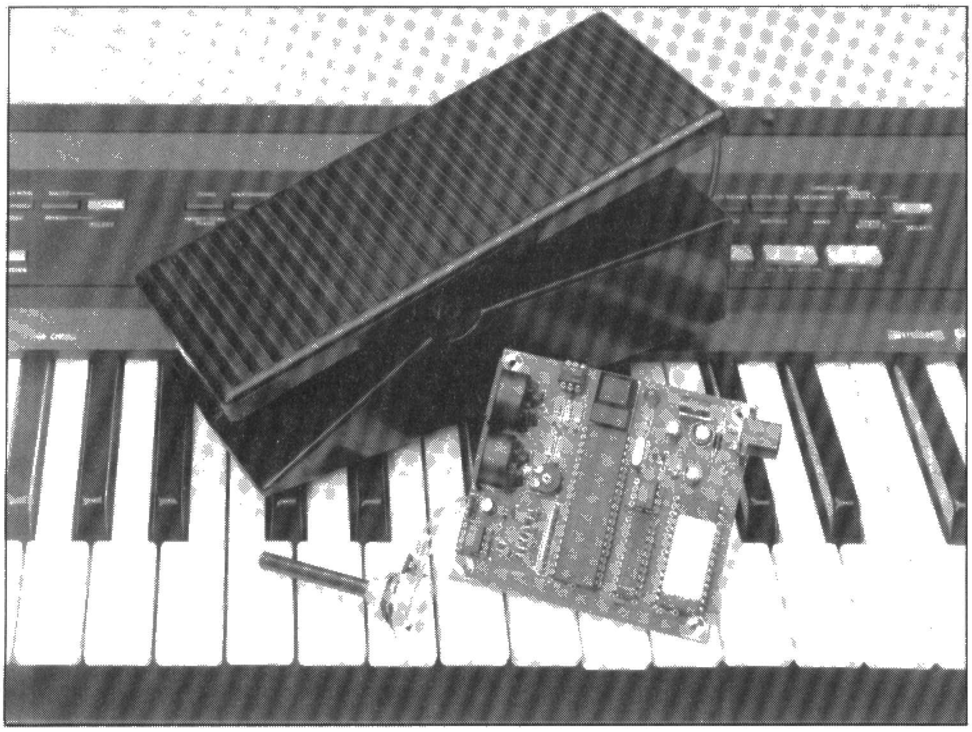
The sound produced by the modified system is a subjective matter. Tests with the prototype showed that some people like the sound with P_1 at maximum resistance (L-R signal), whereas others were inclined to go to the other extreme. Also, it appears that most people liked it better for pop music than for classical music. It may well be that only experimenting with the values of the potentiometer and the electrolytic capacitors gives a sound that is just right for you. [906035]



Good quality medium-frequency or wide-band drive units are perfectly suitable.

MIDI SWELL PEDAL

Design by D. Doepfer



Microcontrollers are intended to make life simpler (?!) and the equipment in which they are used more versatile. The swell pedal presented here uses one and, therefore,can be configured rapidly for a number of functions. Musicians will almost certainly like the design, because it enables a standard swell pedal to be provided with a number of new features for only a small outlay.

Most inexpensive swell pedals on the market are no more than a potentiometer in a box. The potentiometer's resistance varies according to the degree with which the pedal is depressed. The present design is, strictly speaking, an interface between a MIDI system and the existing swell pedal, that is, it provides communication between the pedal and the instrument via MIDI codes. This not only leaves the quality of the sound unaffected, but it also offers new facilities. For instance, the volume can be influenced; the timbre can be adapted; the keying dynamics can be altered and several MIDI channels can be driven simultaneously. Selecting and setting up of the wanted functions and channels remains possible with the use of a standard MIDI keyboard.

Circuit description

A versatile design as outlined above is possible only with the use of a microcontroller, for which a Type 80C32 was chosen—see **Fig. 1**. Since within a MIDI system all communication takes place at a baud rate of 32 kbit s^{-1}, a clock frequency of 12 MHz is used, since the wanted baud rate is easily derived from this.

The microcontroller, IC_1, is linked to EPROM IC_3 via busbuffer IC_2. The application software is stored in IC_3. The buffer is needed to sort out the multiplexed address and data bus.

The level of the ALE (Address Latch Enable) line shows whether the bus carries address signals or data. The ALE signal ensures that the address information is stored in IC_2 at its trailing edge.

The PSEN signal instructs the EPROM to place the data of the selected address on to the databus. After a reset, the controller automatically executes the program in the EPROM.

A discrete 7-bit digital-to-analogue (D-A) converter based on the P_1 ports processes the position of the swell pedal (potentiometer P_2). The 7-bit width enables the digitization of 128 positions of the pedal. This number corresponds to that for the coding in the MIDI protocol.

The output signal of the D-A converter, available across C_4, is compared by IC_4 with the direct voltage at the wiper of P_2. When the output level of IC_4 is high, the D-A signal is lower than the potential at the wiper of the analogue swell pedal. This results in the microcontroller raising the level of the D-A signal, by successive approximation, until the output of IC_4 goes low.

When the comparator changes state, the position of the swell pedal is known and available in digital form. The control range of the D-A converter can be altered to some extent with P_1 to ensure that the 128 steps fall within the (resistance) range of P_2. This arrangement was found necessary because in certain circumstances P_2 did not provide the maximum level of 5 V at the output. If it is found, however, that it does, P_1 may be omitted.

The control program is stored in IC_7, a 2 kbit EEPROM. The programming of this device will be reverted to. For now, it is sufficient to know that this circuit has a 'teach' mode. In this mode, a choice can be made for the function of the pedal and the MIDI channel in which communication takes place. These choices are stored in IC_7 and retained there until new ones are made. The IC has an I^2C interface and communicates with the microcontroller via two I/O lines, P3.6 and P3.7. These lines simulate the communication channels of the I^2C interface.

The address lines of IC_7, A_0, A_1 and A_2, are linked to ground, so that the memory is set to its base address, $A_{0(H)}$.

As already mentioned, IC_7 is programmed in its teach mode. This mode is selected when S_1 is closed, which is indicated by the flashing of D_1.

The MIDI input is via K_1, and the MIDI output via K_2. The input is electrically isolated from the remainder of the circuit by optoisolator IC_6. Such an isolator is present in all MIDI equipment to make it possible for a current loop to be used for MIDI communication.

The incoming 5 V power supply is stabilized by IC_5. Diode D_3 protects the circuit against wrong polarity of the supply lines.

Construction

The circuit is best built on the double-sided printed-circuit board shown in **Fig. 2.** The small size of the board allows it to be built into most swell pedals.

Start by soldering all the passive components into place. Fit IC_3 into an appropriate socket (this makes updating of the application program at a later date a simple matter). All other ICs can be soldered directly on to the board, although sockets are to be preferred.

When the board is populated (including the programmed EPROM), the interface is ready for use.

Use a mains adaptor with an output of not less than 8 V. Since the current drain should not exceed 300 mA in any circumstances, most mains adaptors are suitable.

A small modification needs to be made before the swell pedal can be connected to the interface. Normally, a swell pedal has two connections (it behaves like a variable resistor), but for the present circuit three connections are required. Open the pedal, solder three wires to the potentiometer, and close the pedal again.

If a pedal with an LDR (light-dependent resistor) is used, open the pedal and solder a 5–25 kΩ resistor in series with the LDR. Solder the +ve line to the resistor, ground to the LDR, and use the junction of the resistor and LDR as the 'wiper' of a potentiometer. Close the pedal again.

Connect the interface to the MIDI system via two MIDI cables and switch on the equipment. If all is well, the LED should light twice, the pedal should be active, and MIDI volume commands (MIDI controller #7) should be sent when the pedal is moved.

The range of the pedal should be from 0 to 127; if it is not, adjust P_1 accordingly (which may be done by ear).

If the swell pedal can not be turned to zero, the position of P_2 must be altered until an output voltage of 0 V is obtained. Normally, it will be found possible to adapt the position of the potmeter or the spindle to which it is linked: in practice, it is a matter of a few millimetres at most.

Usage

The interface can be configured entirely to individual requirements. After it has been connected to the system and the mains supply, and this is switched on, it is in the default mode. This means that the volume commands are sent via MIDI channel 1. When the pedal is depressed, the corresponding commands are sent via the interface. The value at which this happens varies between 0 and 127. If the pedal emulates a pitch bender, the code varies between 64 and 127 or between 64 and 0. The pedal can also be used to adapt the velocity and thus the dynamic range of the incoming MIDI data. Note that in this mode data appear at the output only if there are data applied to the input.

If the MIDI equipment does not react to the commands, check the selected channel and ascertain that the polarity of the MIDI cables is correct. Operation of the interface can be checked by pressing push button switch S_1, whereupon the LED should begin to flash.

It is also possible that the expander used does not react to the volume command. There are some expanders, such as the Yamaha EMT10, which do not recognize this command. There are also expanders that use the command only for generating an after-tone. This means that the volume of an already generated tone can not be influenced.

After receiving a program change command, the interface always sends an instruction containing the volume setting. This is necessary because some expanders switch to maximum volume (127) upon receiving a program change command.

Teach mode

The default mode, at which the pedal functions as MIDI controller #7 and commands are sent in MIDI channel 1, can be altered by setting the interface to the teach mode. This is indicated by the flashing of the LED. The teach mode is used to select the MIDI channel at which the interface is required to work. This can be done with a standard MIDI keyboard when the interface input is linked to the MIDI output of the keyboard.

Key	**MIDI channel**
36 (C)	1 on
37 (C#)	2 on
38 (D)	3 on
39 (D#)	4 on
40 (E)	5 on
41 (F)	6 on
42 (F#)	7 on
43 (G)	8 on
44 (G#)	9 on

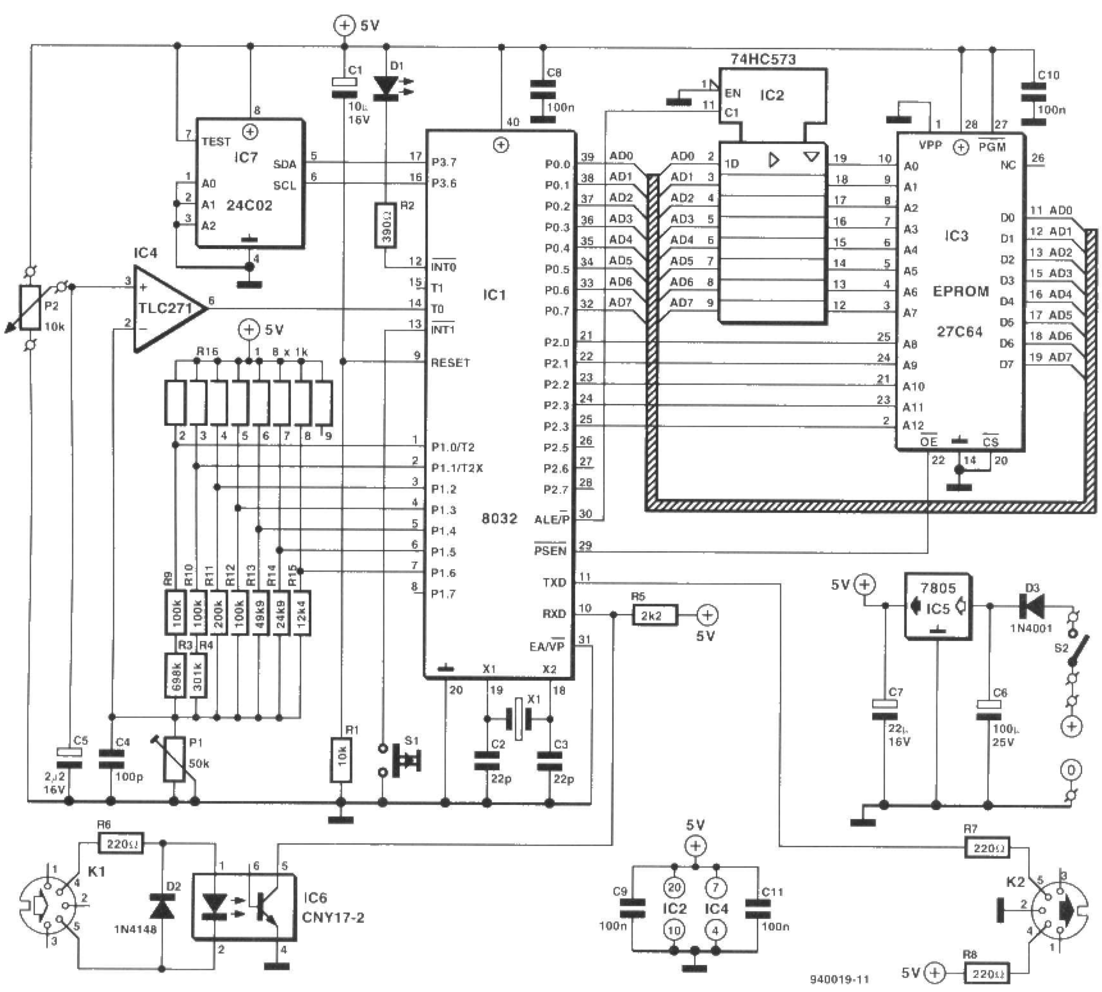


Fig. 1. Circuit diagram of the interface for the MIDI swell pedal.

45 (A)	10 on
46 (A#)	11 on
47 (B)	12 on
48 (C)	13 on
49 (C#)	14 on
50 (D)	15 on
51 (D#)	16 on
60 (C)	1 off
61 (C#)	2 off
62 (D)	3 off
63 (D#)	4 off
64 (E)	5 off
65 (F)	6 off
66 (F#)	7 off
67 (G)	8 off
68 (G#)	9 off
69 (A)	10 off
70 (A#)	11 off
71 (B)	12 off
72 (C)	13 off
73 (C#)	14 off
74 (D)	15 off
75 (D#)	16 off

The wanted command is sent by depressing the relevant key on the keyboard when the LED flashes. The dynamic range properties sent in the MIDI command are not used by the interface. To keep the system easy to use, the note C is chosen as the starting point for both the switch on and the switch off commands.

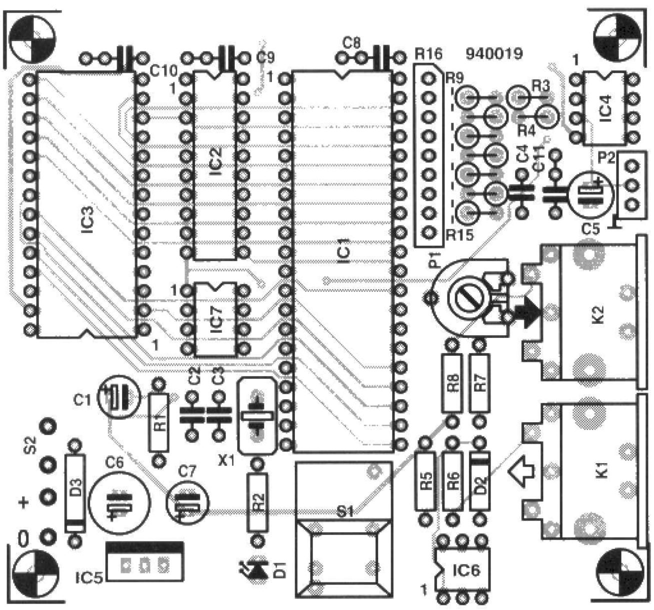
It may of course happen that it is no longer clear which of the channels are active. In that case, there is only one solution: switch off all channels and switch on the wanted ones.

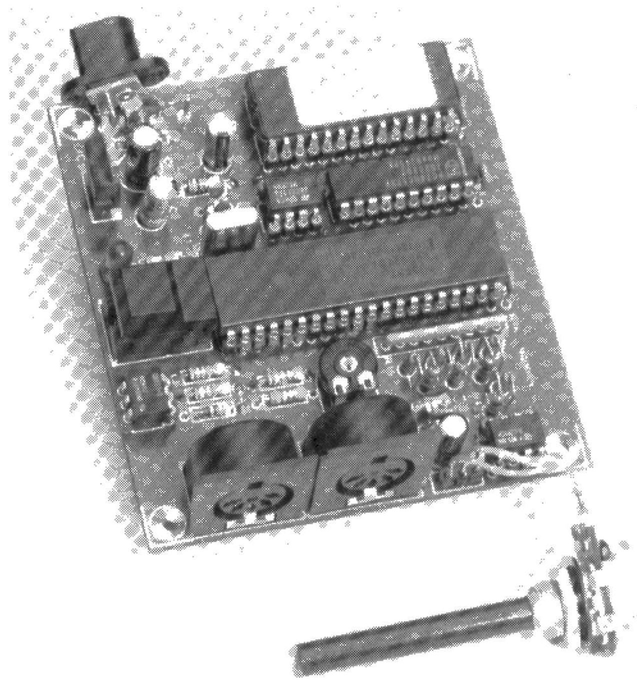
After all wanted channels have been selected, the teach mode is discontinued when the push button on the interface is pressed or when a program change command is sent. The LED then stops flashing and the setting is stored in the EEPROM until the teach mode is selected again.

Other functions

So far, the interface has been used to modify a simple swell pedal into a modern digital pedal to influence the volume of the sounds. However, this is the standard configuration: the interface offers other functions, such as velocity control, after-touch or pitch bending.

Program Number	Function
1	volume (controller #7)
2	modulation (controller #1)
3	portamento (controller #2)
4	free choice
5	after-touch (mono)
6	pitch bend (entire range)





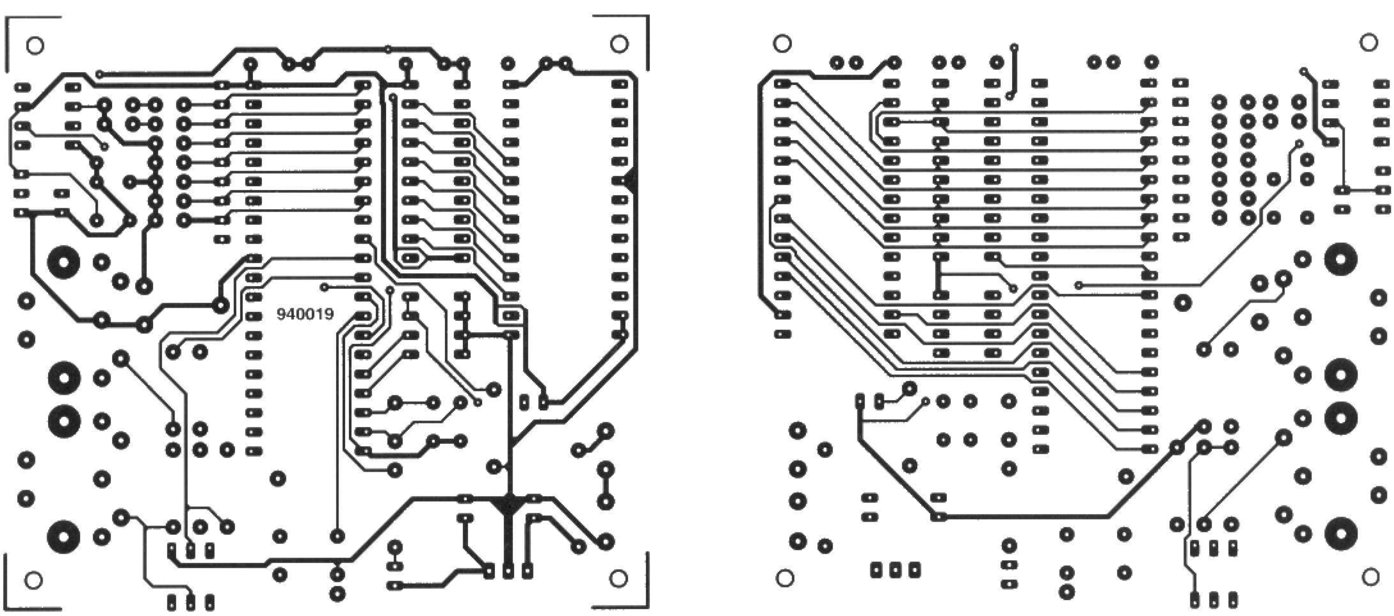


Fig. 2. The double-side printed-circuit board for the interface.

7	pitch bend (positive)
8	pitch bend (negative)
9	velocity (touch dynamics)
16	data speed

The interface is set to the previously defined controller numbers (#7 for volume; #1 for modulation; #2 for portamento) with the aid of program change commands 1, 2 and 3. A different controller number, if desired, can be set with program change command 4. With this command, the interface adopts the controller number that was sent prior to the program change command. If, therefore, a random controller number is desired, a MIDI controller command with that number must be sent to the interface followed by program change command 4.

Program numbers 6, 7 and 8 may be used to imitate the various functions of a pitch bender. Program number 6 provides the range 0–127 with a neutral zone around number 64: this simplifies the setting of a neutral position. Program numbers 7 and 8 simulate the positive swing (64–127) and negative swing (64–0) of the pitch bender respectively.

The velocity function is a special one that can be selected with number 9. If then the pedal is depressed, it will no longer cause MIDI data to be sent but data that are being sent on the selected MIDI channel to be adapted. The velocity value contained in the MIDI command is then adapted in line with the position of the swell pedal. In the highest position of the pedal, the velocity value is multiplied by 1, that is, it remains unchanged. The more the pedal is depressed, the smaller the factor with which the value will be multiplied. This increases the touch dynamic range.

This function was found useful because there are MIDI instruments on the market (expanders, keyboards, and others) that do not support the MIDI volume command (controller #7). In spite of that deficiency, the interface makes it possible to influence the sound of such instruments. Also, keyboards that have no touch-sensitive keys may benefit from this function.

Note that the velocity function affects only the dynamic range of the note-on command; the note-off instruction retains its previous value.

Program number 16 influences the frequency (speed) at which commands are transferred through the interface. This function is provided because some instruments, owing to less-than-perfect software, suffer from timing errors if the frequency is too high.

With the interface in the teach mode, it will be noticed that the frequency with which the LED flashes changes when the pedal is depressed. The flashing rate is an indication of the speed at which the commands are transferred. Sometimes it may appear as if the LED lights continuously. This is because the human eye can not follow the high frequency flashing.

Set the frequency with the pedal and give a program change command 16. This setting is stored in the EEPROM. The default frequency is the lowest: about 15 Hz. This means that 15 commands are sent every second. The precise value of the highest positions depends on several factors. Since each MIDI volume command consists of three instructions, 45 bytes per second are sent at the lowest frequency. If 16 channels are active, 16×45=720 bytes per second are sent. Therefore, the ultimate data rate depends on the sampling frequency and on the number of active MIDI channels. The maximum sampling frequency is, therefore, limited by the number of active channels. The optimum setting can be found by trial and error only.

Parts list

Resistors:
R_1 = 10 kΩ
R_2 = 390 Ω
R_3 = 698 kΩ
R_4 = 301 kΩ
R_5 = 2.2 kΩ
R_6, R_7, R_8 = 220 Ω
R_9, R_{10}, R_{12} = 100 kΩ
R_{11} = 200 kΩ
R_{13} = 49.9 kΩ
R_{14} = 24.9 kΩ
R_{15} = 12.4 kΩ
R_{16} = 8×1 kΩ array
P_1 = 50 kΩ (47 kΩ) preset potmeter
P_2 = 10 kΩ (potmeter in pedal)

Capacitors:
C_1 = 10 μF, 16 V
C_2, C_3 = 22 pF
C_4 = 100 pF
C_5 = 2.2 μF, 16 V
C_6 = 100 μF, 25 V
C_7 = 22 μF, 16 V
C_8–C_{11} = 100 nF

Semiconductors:
D_1 = LED
D_2 = 1N4148
D_3 = 1N4001

Integrated circuits:
IC_1 = 80C32
IC_2 = 74HC573
IC_3 = 27C64 (p. 110 – Ref. 946635-1)
IC_4 = TLC271
IC_5 = 7805
IC_6 = CNY17
IC_7 = 24C02

Miscellaneous:
K_1, K_2 = 5-pin DIN socket, 180°
S_1 = push-button switch with single-pole make contact
S_2 = single-pole, single-throw switch
X_1 = 12 MHz crystal
PCB Ref. 940019 (p. 110)

[940019]

PIC® PROGRAMMING COURSE

PART 1: INTRODUCTION

Large designs traditionally required complex and extensive digital circuits are now simple to realize by virtue of the small, powerful, PIC® microcontroller developed by Microchip Technology. The main application areas of PICs are automotive electronics, machine controls, and test and measurement equipment. Particularly logic circuits based on time or count pulses often become quite complex if ordinary logic components are used. A single PIC however can do the same at a smaller outlay, offering a staggering degree of circuit simplification. Microchip's PIC16C5x family is a class of its own in microcontroller land, and forms the subject of the present short course. Apart from programming aspects, the hardware will also receive some attention.

PIC = Peripheral Interface Controller, a registered trademark of Microchip Technology, Inc.

By our editorial staff.

Source: Microchip Data Book, Microchip Technology Inc.

THE PIC16C5x family is a product of Microchip Technology Inc., of Chandler, Arizona, in the United States of America. The family consists of a series of CMOS microcontrollers featuring internal data and program memories. The program memory has a word size of 12 bits, which is obviously more than that of competitive 8-bit controllers. As indicated in the 'portrait of a family' inset in our earlier article describing a PIC programmer (Ref. 1), the size of the program and data memory depends on the type of controller.

The fully static design of the microcontrollers allows the clock frequency to be reduced to d.c. The advantage of the 12-bit word size is that most instructions require only one 'word', inclusive of their operand. The controller has 33 easy to remember instructions. With the exception of jumps (or 'branches'), these take one machine cycle. Consequently, Microchip advertises the PIC16C5x as a device which employs a RISC-like architecture (RISC = reduced instruction set computer), which is marked by a compact but fast sequence of instructions, each of which is executed in one machine cycle.

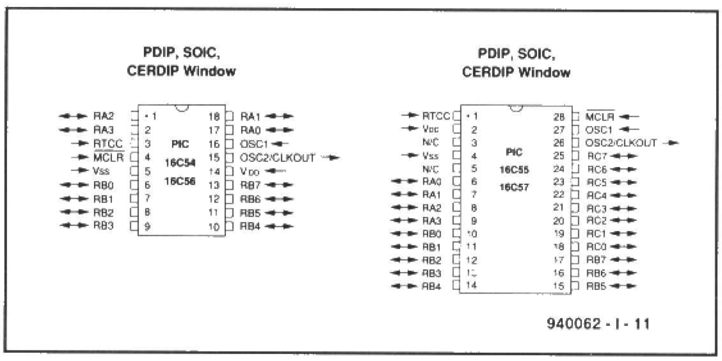


Fig. 1. Depending on the device, the 'window' type PIC microcontrollers come in an 18-pin or a 28-pin case.

Apart from the differences in memory size, the controllers in the 16C5x family have different numbers of I/O lines available, and different types of clock oscillator. Furthermore, the devices come in different enclosures, two of which are shown in **Fig. 1**. Comparing the different cases, the most conspicuous feature is the presence or absence of a glass window. The windowless, and therefore cheaper, OTP (one-time programmable) versions is suitable for high-volume production. The window versions contain an EPROM, and are ideal for code development, since their program memory can be erased using ultra-violet light. All devices in the 16C5x family offer a copy protection bit. When set, this bit makes it impossible to read code from a programmed controller.

Internal hardware

The PIC architecture is based on a register file concept with separate bus and memories for data and instructions. This so-called Harvard architecture enables the processor to execute one instruction and at the same time fetch the next from the program memory. Unfortunately this 'look-ahead' mode can not be used for a number of branch instructions. Consequently, these instructions take two machine cycles instead of one.

The controller architecture is shown in **Fig. 2**. The program memory and the associated counter, the instruction register and the instruction decoder are found in the upper left-hand corner of the diagram. Below these elements sits the ALU (arithmetic/logic unit) with its working register, W. Central to the right is the data memory (general-purpose register file), and above that, the clock generator with watchdog, and the I/O register.

Program memory

The program memory consists of 'pages' of 512 words each. The PIC16C54 and 16C55 have one such page, while the PIC16C56 has two. Address bit A9 is used to select between the two pages. The PIC16C57 has four pages, which are selected with the aid of bits A9 and A10.

The structure of the program memory is illustrated in **Fig. 3**. Sequencing of microinstructions is controlled via the program counter (PC) using a 9-bit or 11-bit wide address. Some precautions have to be taken when executing branch instructions, or starting a subroutine. Problems may then occur if it is assumed that the data is 8 bit wide, which makes it impossible to fit in operands having a width of 9 bits or more.

The effect of a program instruction on the address bits of the program counter is illustrated in **Table 1**. The

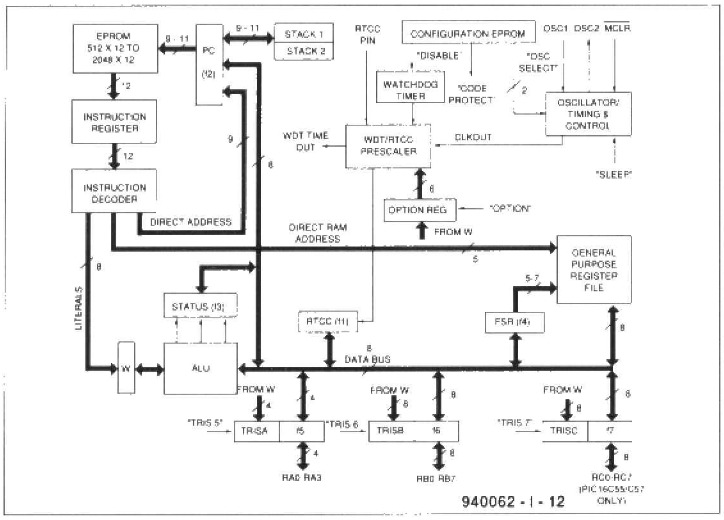


Fig. 2. General architecture of a PIC16C5x microcontroller. Note the different widths of the internal buses.

programmer should take the above restriction as regards word size into account. With the exception of one GOTO instruction, instructions which alter the program flow may only be used in the lower half of the memory block (A8=0). Furthermore, bits 5 and 6 of the f 3 register must be correctly programmed when the PIC16C56 or 16C57 is used.

These bit are not automatically modified when the software jumps between pages. Making sure that they are modified appropriately is the task of the programmer.

Data memory and register file

Microchip Technology calls the PIC's internal memory 'register file' (RF), because a register is a memory location which can be accessed directly by the ALU. The set of the registers is then called a file.

The structure of the register file is shown in **Fig. 4**. Registers fall into the categories 'operational', 'I/O' or 'general-purpose'. The operational registers not only serve to control the core of the PIC, but also make results of certain actions available to the program. The I/O registers are used to gain access to the ports. Finally, the general-purpose registers are used to store data. To the user, all registers behave in the same way.

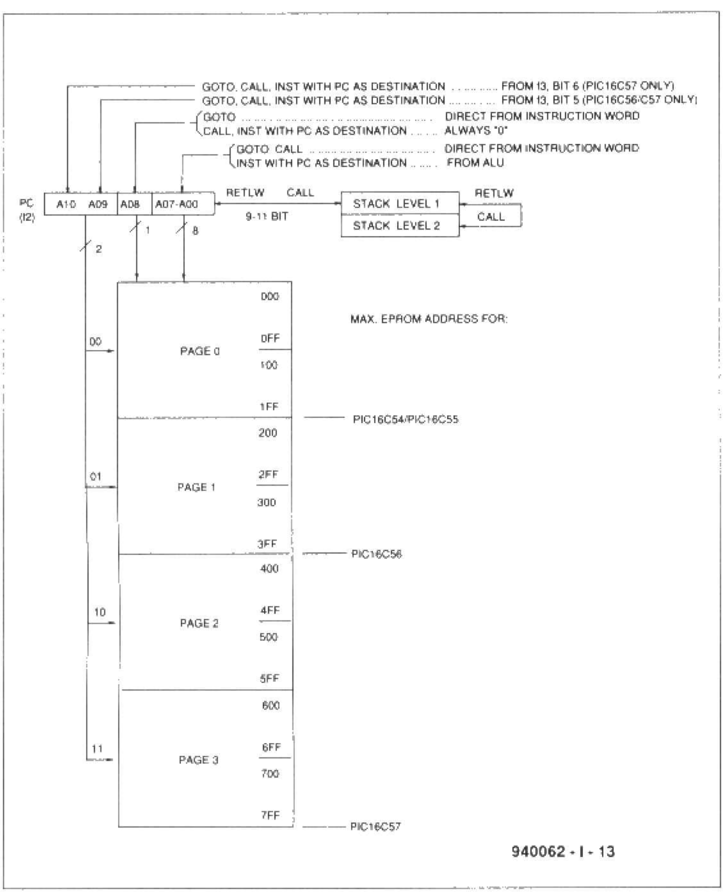


Fig. 3. Structure of the program memory, which is divided into pages.

Operational registers

Register f0, Indirect Data Addressing

This register give the programmer indirect access to the data. Access is gained in conjunction with the FSR register described further on. A read or write command to register f0 is treated such by the controller that the ALU selects the location of which the address is contained in the FSR register. The combination of register f0 and the FSR register creates ways to manipulate memory areas in a very efficient way. In the (trivial) case of f0 being read through indirect addressing (FSR then contains the value 00H), the value 00H is read. If f0 is written to via indirect addressing, the result is a NOP.

Register f1, RTCC (Real-time Clock/Counter register)

This register may be compared to a location in the data memory. The content of this location is continuously incremented using a clock signal. The clock signal may be external, and applied to the RTCC input, or internal, when it is derived from the controller's instruction cycle clock. An internal prescaler enables the clock signal to be scaled as required. Further details on this feature are included in the discussion of the watchdog timer.

PIC16C5x PROGRAMMING

This short course is aimed at providing an introduction into programming and hardware aspects of the PIC16C5x family of microcontrollers manufactured by Microchip Technology Inc.

An **assembler** will be offered on disk later in the course. This assembler is distributed with the permission of Microchip tecnology Inc., and supports the PIC16C5x and PIC16Cxx series of controllers. It offers a full featured macro and conditional assembly capacity. It can also generate various object code formats including several hex formats to support Microchip's proprietary devleopment tools as well as third party tools. Also supported are hex (default), decimal and octal source and listing formats. An assembler users manual is available from Microchip technology distributors for detailed support.

The disk, which will be distributed via the *Elektor Electronics* Readers' Services will also contain a software **simulator**.

The files produced by the assembler can be downloaded to the **PIC programmer** described in *Elektor Electronics* March 1994.

JOIN THE COURSE!

Register f2, PC (Program Counter)

The program counter generates the addresses of the locations in the program memory. Depending on the device type, the program counter and its associated two-level hardware stack is 9, 10 or 11-bit wide, using bit A8, A9 and A10 as extensions of A0-A7. **Table 2** summarizes the options.

The PC is normally incremented by one after an in-

PC address bit	Goto	Call	PC instruction
A10	bit 6 of f3	bit 6 of f3	bit 6 of f3
A9	bit 5 van f3	bit 5 of f3	bit 5 of f3
A8	from instr.	always 0	always 0
A7-A0	from instr.	from instr.	from ALU

Table 1. The effect of the different instructions on the address bits.

Device	PC and stack width
PIC16C54	9 bit (A8)
PIC16C55	9 bit (A8)
PIC16C56	10 bit (A8,A9)
PIC16C57	11 bit (A8,A9,A10)

Table 2. The width of the program counter and the associated hardware stack.

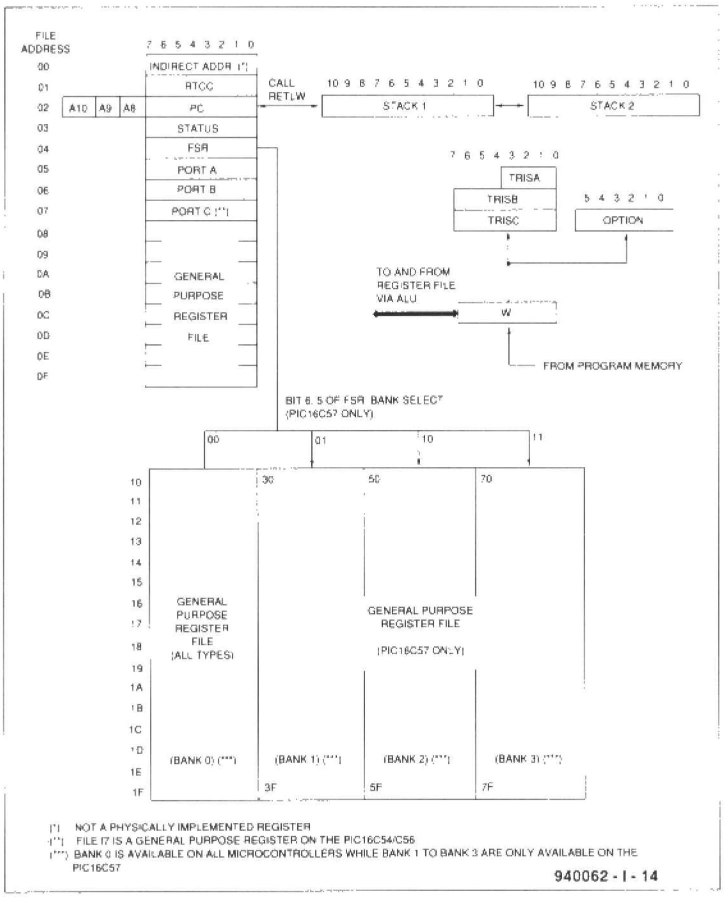


Fig. 4. The structure of the data memory is fairly complex because it is divided into banks. A key role is played by the FSR register.

struction has been executed. There are, however, a number of instructions which follow a different pattern:

GOTO
Bits A0-A8 are loaded directly, while A9 and A10 are fetched via bits 5 and 6 of the status register (PA0, PA1). When using the PIC16C56 and 16C57, the status register must be loaded appropriately.

CALL
The CALL instruction is different from GOTO in that bit 8 is always cleared to 0. Consequently the range of this instruction is limited.

All other write instructions to the PC operate in the same way as the CALL instruction, i.e., they clear A8, and read A9 and A10 from the status register (16C56/57 only).

Register f3, Status Word Register
The C, D and Z flags (bits 0, 1 and 2) in this register provide the status of an arithmetic operation carried out by the ALU. The PD flag (bit 3) and the TO flag (bit 4) indicate the reset status, while PA0 (bit 5) and PA1 (bit 6) are used as auxiliary bits (A9 and A10) for certain program counter operations (CALL, GOTO, see above).

Register f3

bit 7	bit 6	bit 5	bit 4	bit 3	bit 2	bit 1	bit 0
PA2	PA1	PA0	TO	PD	Z	DC	C

Bit	Function
PA2	general purpose R/W, reserved for future use*
PA1	page preselect bits
PA0	PIC 16C54 and PIC16C55 PA1 and PA0: general purpose read/write* PIC 16C56 PA0: 0= bank 0 (000H-1FFH) 1= bank 1 (200H-3FFH) PA1: general purpose read/write* PIC 16C57 PA1/PA0: 00 = bank 0 (000H-1FFH) 01 = bank 1 (200H-3FFH) 10 = bank 2 (400H-5FFH) 11 = bank 3 (600H-7FFH)
TO	Time-Out bit Set to 1 during power-up and by the CLRWRT and SLEEP commands. Reset to 0 by a watchdog timer time-out. Not affected by other commands.
PD	Power-Down bit Set to 1 during power-up and by the CLRWRT command. Reset to 0 by a SLEEP command. Not affected by other commands.
Z	Zero bit Set if the result of an arithmetic logic operation is 00H.
DC	Digit Carry/Borrow bit Carry/Borrow bit (bit 3) of result of ADDWF and SUBWF instructions.
C	Carry/Borrow bit Carry/Borrow bit (bit 7) of result of ADDWF and SUBWF instructions. Also used with RRF and RLF instructions.

* bits must not be used if application is to run on future systems.

Event	TO	PD
Power-up	1	1
WDT timeout	0	x
Sleep instr.	1	0
CLRWDT instr.	1	1

x = don't care
WDT = watchdog-timer

Table 4. Status of the TO and PD bits as a function of different functions.

TO	PD	Reset caused by:
0	0	WDT wake-up from SLEEP
0	1	WDT timeout (not during SLEEP)
1	0	MCLR wake-up from SLEEP
1	1	Power-up
x	x	low pulse on MCLR input.

State of TO and PD not affected (x) until an event from Table 4 occurs.

Table 5. TO and PD status after a reset.

Bits 3 and 4 (TO and PD) in the status word register are not affected by an 8-bit wide write action.

If the register is used to indicate the result of an arithmetic operation, it should be noted that the status bits are set after the following write.

It is recommended to use only the BCF, BSF and MOVWF instructions to alter the status register, since these do not affect any status bit.

Register f4, FSR (File Select Register)
PIC16C54/55/56: bits 0 trough 4 select one of the 32 available file registers in indirect addressing mode, i.e., using register f0. Bits 5, 6 and 7 are read-only, and always at 1.

If no indirect addressing is used, the FSR may be used as a 5-bit wide general-purpose register.

PIC16C57 only: bits 5 and 6 select the current data memory bank, both in indirect and direct addressing modes. Note that this is only valid for the register addresses 10H through 1FH, since addresses 0H through 0FH always point to the same registers. Finally, bit 7 is always at 1.

(940062-1)

Continued in the September 1994 issue.

Reference:

1. PIC programmer. *Elektor Electronics* March 1994.

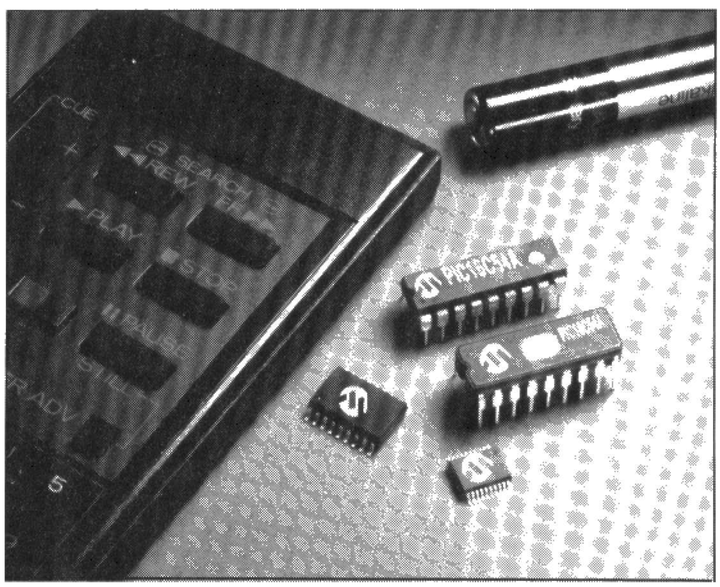
MICROCHIP ANNOUNCES FURTHER IMPROVEMENTS IN MICROCONTROLLERS

Microchip's PIC16C54 8-bit microcontroller is now available with reduced power consumption, improved electrical characteristics and denser packaging. Using an advanced 0.9-micron double-layer metal wafer fabrication process, the new one-time programmable PIC16C54A can be powered from a single lithium-ion battery, making it an ideal solution for portable aplications such as pagers and remote controls.

The PIC16C54A is the first of the PIC16C5x family to be manufactured using the 0.9-micron process. The high-speed RISC-like 8-bit device operates at up to 20 MHz, and provides faster instruction execution than any other 8-bit microcontroller in its price range. An on-chip EPROM fuse configurator allows designers to select on-chip R/C timing circuits and crystal/resonator options to reduce component count, cost and board space requirements. On-chip memory facilities include 512 words of EPROM for program storage and 25 bytes of static RAM for data. On-chip peripherrals include an 8-bit realtime clock/counter with programmable prescaler, a watchdog timer, and 12 I/O lines with individual directional control. The PIC16C54A operates between 2.5 and 6.0 volts, and includes a power-down or sleep mode which reduces current drain to less than 4 μA when executed.

The new PIC16C54A is available in plastic DIP and SOIC, and also in a compact SSOP configuration.

Further information from



Microchip Technology Inc., 2355 West Chandler Blvd., Chandler, AZ 85224-6199, USA. Tel. 602 786-7200. Fax: 602 899-9210.

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Arizona Microchip Technology Ltd., Unit 3, The Courtyard, Meadowbank, Furlong Road, Bourne End, Bucks SL8 5AJ. Tel.: (0628) 850303. Fax: (0628) 850178.

UK distributors:
Farnell Electronics, tel. (0532)792715, fax: (0532) 63340.
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Polar Electronics PLC, tel.: (0525) 377093, fax: (0525) 378367.

Sweden:
MEMEC Scandinavia AB, tel.: 08 6434190, fax: 08 6431195.

Denmark:
Exatec A/S, tel. 044 92 7000, fax: 044 92 6020.

SOFTWARE EMULATION OF RC5 INFRA-RED CODE

A set of programs is described that allows a PC to send infra-red remote control commands to a TV set, video recorder or hi-fi stereo chain, using the RC5 standard. As regards hardware, all you need is one infra-red transmitter LED.

Design by M. Claessen

ONE very obvious trend in audio/visual consumer electronics is the use of remote controls. Designed to maximize user comfort, infra-red remote controls enable nearly all functions of audio/video equipment to be selected and controlled from 'armchair distance'. This article shows that remote control is not only a 'luxury', but also makes it easy to set up links with other equipment, say, a PC.

Using suitable software, a video recorder, TV set or stereo rack can be controlled by a personal computer (PC) without having to change a thing on this equipment. The hardware extension required at the PC side of the link is, indeed, minimal, consisting of a single infra-red light emitting diode (IR LED), which is connected in series with the loudspeaker found in any PC. A small program then ensures that the right RC5 commands are generated and transmitted by the LED. The loudspeaker will function as before.

RC5 is a system developed by Philips for communication using infra-red light. The system is also used by a number of other manufacturers of audio/video equipment. For the present project the designer has based the software on the RC5 standard and code set. The source code contained on disk should enable experienced programmers to adapt the program so that it can be used for, say, Sony or Pioneer equipment.

Codes for frequently used addresses and commands

System address	Apparatus
0	TV
2	Teletext
5	Video recorder
7	Experimental
16	Preamplifier
17	Receiver/tuner
18	Tape/cassette recorder
19	Experimental
20	CD player

Command number	Command
0-9	0-9
12	Standby
13	Mute
14	Presets
16	Volume +
17	Volume –
18	Brightness +
19	Brightness –
20	Colour saturation +
21	Colour saturation –
22	Bass +
23	Bass –
24	Treble +
25	Treble –
26	Balance right
27	Balance left
48	Pause
50	Fast reverse
52	Fast forward
53	Play
54	Stop
55	Record

The program

The function of the program is to translate instructions or commands into a pulse code signal that can be applied to the loudspeaker connection on the PC's motherboard. This loudspeaker output is fairly rudimentary, and usually consists of an 'power' inverter only, which drives a miniature loudspeaker directly. The maximum power delivered to the speaker is about 150 mW .

In a normal infra-red remote control link, the encoding of selected commands is performed by a simple integrated circuit contained in the handheld unit. In the present case, the encoding function is assumed by a program running on the PC. Since generating the digital code is time-critical, a machine-language program is used for that purpose. That allows the program, ELEKFUNC.EXE, to be called by routines written in 'higher' languages, for instance, Turbo Pascal.

MAIN SPECIFICATIONS

Code used:	RC5
Range:	5 to 13 m
PC requirements:	MS-DOS
PC type:	XT/AT/386/486
Source code:	on disk
Emulation:	Philips RC5903
IR diode:	LD271 or LD274
Transmit power:	max. 150 mW

Apart from ELEKFUNC.EXE, the diskette supplied through the Readers Services (order code **1901**) contains two more programs: SWITCH.EXE and RC5.EXE. The latter enables you to select and transmit RC5 commands in a simple manner. SWITCH.EXE may be used to transmit a system address and a key code 'manually', for instance, 'SWITCH 24 35'.

The diskette also contains the source code of all three programs (Pascal for SWITCH and RC5, and assembler for ELEKFUNC). The source code files should enable you to fine-tune the software to personal requirements. In this context, **Ref. 1** should be mentioned, since that design enables you to set up a complete wireless control system capable of switching apparatus in the home on and off. One application would be a lighting control system managed by the PC while you

are away on holiday. One condition for such a system to work is, of course, that the transmitter diode connected to the PC 'sees' the receiver diode in the apparatus to be controlled.

The hardware

As already indicated, an infra-red transmitter LED must be fitted inside the PC. In most PCs, the loudspeaker is connected to the motherboard via a thin cable and a 4-way connector. Usually, only three of the four wires are used. The fourth connector pin serves to polarize the connector so that it can not be fitted the wrong way around. As far as we have been able to ascertain, the pin functions on this connector are as follows: pin 1 is data, pin 2 is not connected, pin 3 is ground, and pin 4 is +5 V. Connect the IR LED in series with the loudspeaker, making sure that the cathode goes to the PC ground connection. Alternatively, the diode may be connected across (i.e., in parallel with) the loudspeaker. In particular when the loudspeaker (or buzzer) is fitted on the motherboard, that may be the only viable option. (930097)

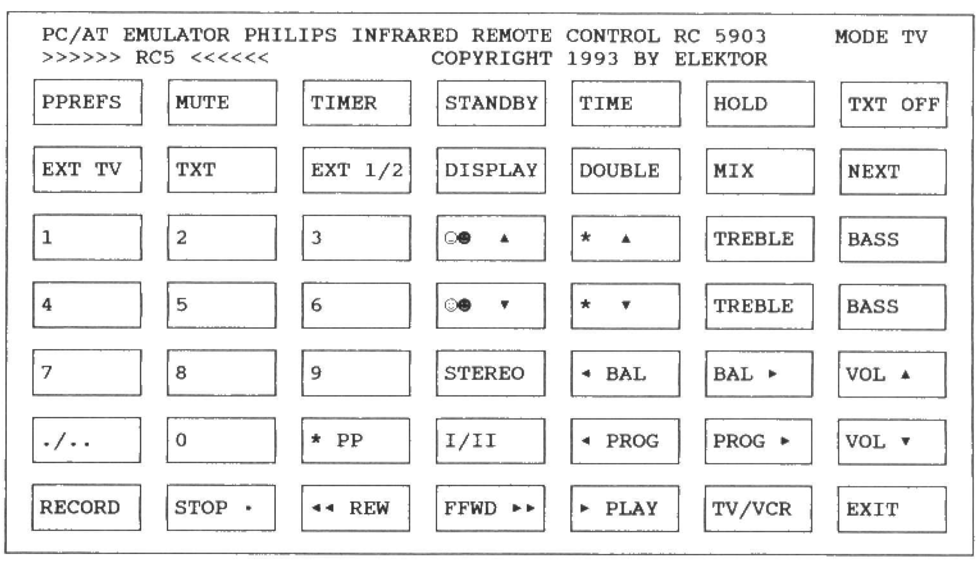


Fig. 1. Screendump showing how the program RC5.EXE enables an infra-red remote control to be simulated on an MS-DOS PC.

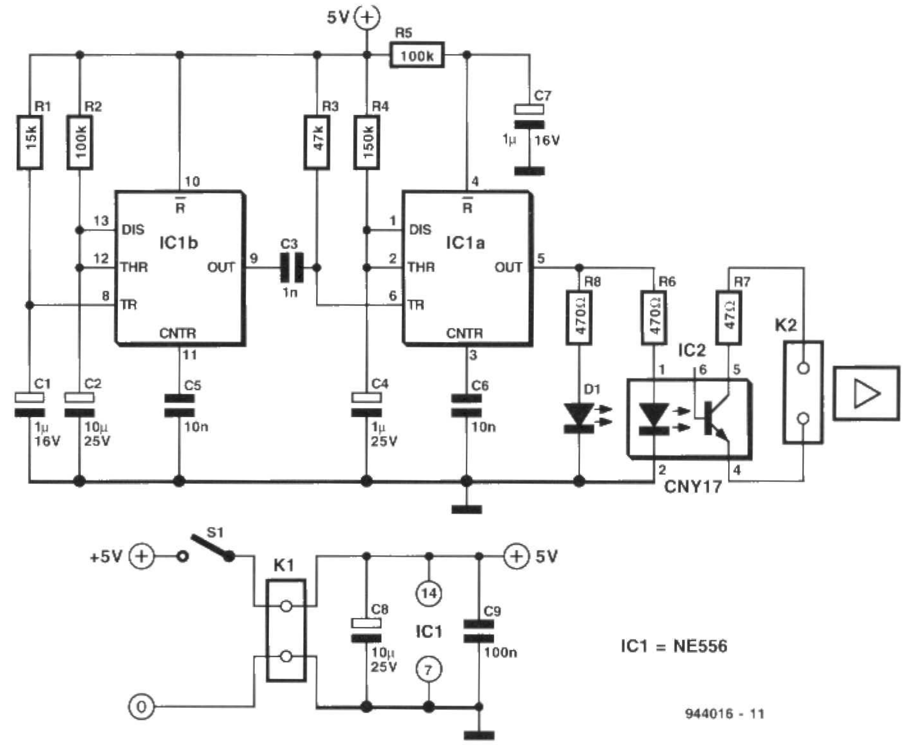
Reference:

1. Universal RC5 code infra-red receiver. *Elektor Electronics* January 1992.

AUTO-PLAY FOR CD PLAYER

Many CD players are equipped with a timer, which starts the player automatically as soon as the mains is switched on (perhaps by a time switch). There are, however, several makes that do not have this facility: the present circuit is intended for these.

The circuit is based on a dual timer Type NE556. The 5 V supply is derived from the CD player. When switch S_1 is closed and the mains is switched on to the CD player, the present circuit will be powered via connector K_1.



Immediately upon the supply voltage becoming available, IC_{1b} receives a trigger signal via network R_1-C_1. The timer then provides a pulse of about 1.2 s (time constant R_2-C_2). During this short period, the CD player is initialized. After the pulse has decayed, IC_{1b} triggers the other timer, which in its turn switches on the transistor in optoisolator IC_2 for 0.2 s (time constant R_4-C_4). The transistor is connected in parallel with the contacts of the play switch on the D player, so that this starts playing the CD in the disc compartment.

Since the switch circuit of most CD players is multiplexed one way or another, the optoisolator is imperative as it prevents the other signals being affected or even short-circuited to earth.

Since the transistor in the optoisolator conducts in only one direction, wiring it up correctly to the play switch must be done by trial and error. In other words, if it does not work, interchange the two wires from the transistor to the play switch. It is also possible to operate a reed relay via IC_2 to short-circuit the contacts of the play switch.

Light-emitting diode D_1 serves as pulse indicator. It lights the moment IC_{1a} generates a pulse. The diode is not essential and, together with R_8, may be omitted.

Use an NE556, not a TLC556, since this does not provide enough current to drive two LEDs.

The circuit should preferably be fitted inside the CD player, with switch S_1 mounted on the front or rear panel.

The 5 V line is taken from the 5 V power supply in the player. The circuit draws a current of not more than 40 mA.

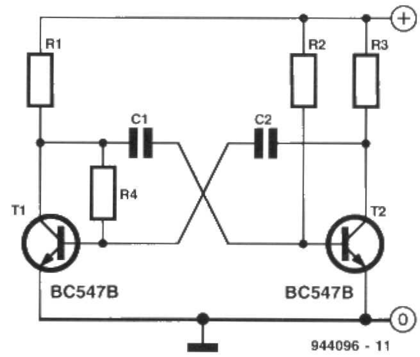
Design: G. Renker
[944016]

SELF-STARTING MULTIVIBRATOR

It is not too well-known that if the characteristics of the two transistors making up a multivibrator are near-identical, it is very difficult to get the multivibrator started.

The present circuit shows how a small modification can make the basic cicuit sufficiently asymmetrical to ensure that he multivibrator always starts unfailingly. The only drawback is that the duty factor is affected slightly, but in practice this will normally not be very important.

In the diagram, resistor R_4 connects the base and collector of T_1, so that this transistor can no longer become saturated. Instead, it works as a self-setting amplifier. A consequence of this is that the amplified noise on the collector of T_2 also appears on the collector of T_1. Because of this coupling, T_2 is provided afresh with the amplified noise signal via C_1. This ensures tat the noise level increases rapidly so that the multivibrator starts.



Putting values to the circuit elements is straightforward. Take a collector current of 1–10 mA. If it is 1 mA, the value of R_1 and R_3 is equal to the supply voltage times 1000. The value of R_1 and R_4 should be equal to a quarter of the amplification factor of the transistors, multiplied by the value of R_1. The value of C_1 and C_2 is the same and is calculated from

$$C = f / (1.4 \times R_3).$$

Design: C. Clarkson
[944096]

PROGRAMMABLE AMPLIFIER

The principle of the amplifier is simple: take a common-or-garden operational amplifier and make the feedback loop switchable with the aid of a multiplexer. Supplying the multiplexer with a 3-bit binary word enables one of up to six different amplification factors to be selected.

In the diagram, IC_{1b} is the amplifier stage proper. Its feedback loop is split into a number of discrete resistors, R_4–R_{10}, which can be switched by multiplxer IC_2. With the specified values, there is a choice of six gains: +20 dB; +10 dB; 0 dB; –10 dB; –20 dB; and –30 dB.

The amplifier/multiplexer is preceded by an input buffer, IC_{1a}. The input is protected against overload by R_1, D_1 and D_2. Conversely, the high input impedance of IC_{1a} obviates overloading of the output of a connected apparatus by the present circuit.

Although the channel resistance of the multiplexer is fairly high at 220 Ω compared with the values of R_4–R_{10}, this has no detrimental effects since it is in series with the relatively high input impedance of IC_{1b}.

The capacitances of the discrete analogue switches in the multiplexer have some adverse effect on the signal, but this stays well within acceptable limits. Up to 100 kHz, the circuit performs well; only above this frequency does a sine wave begin to resemble a triangle.

Harmonic distortion is ≤0.001% at 1 kHz and ≤0.01% at 20 kHz. With the gain set to +20 dB, the signal-to-noise ratio is ≥95 dB referred to 1 V with the input short-circuited.

Te rather unusual supply voltage of ±16.5 V was chosen deliberately to obtain a maximum input voltage of 10 V r.m.s. with gains of –20 dB and –30 dB. If this is considered not very important, the supply may be lowered to ±15 V.

The circuit draws currents of +18 mA and –9 mA.

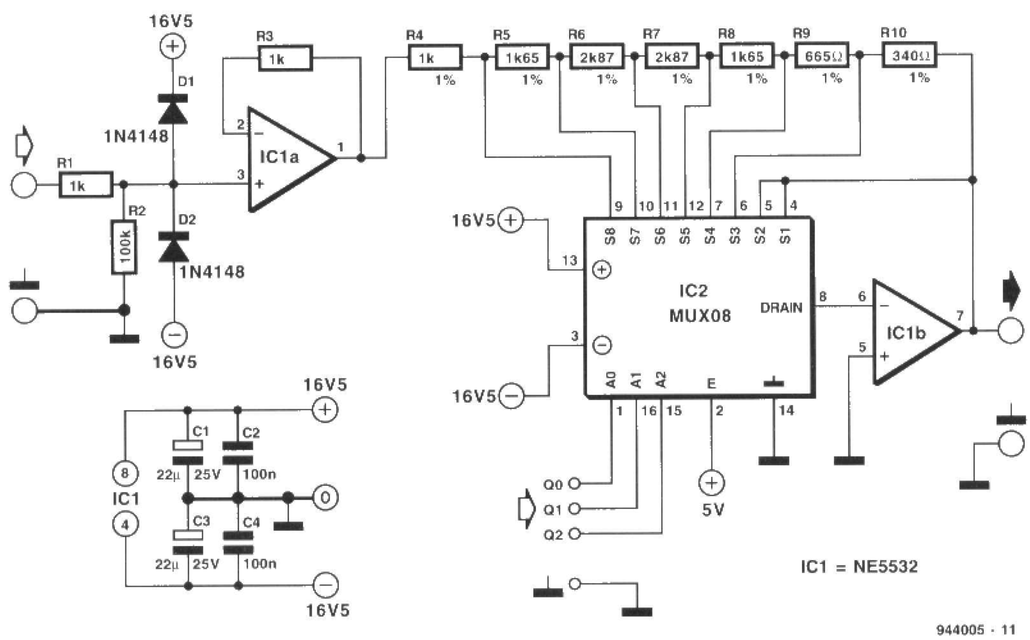
The relationship between the binary word and the resultant gain is:

Q2	Q1	Q0	gain
1	1	1	+20 dB
1	1	0	+10 dB
1	0	1	0 dB
1	0	0	–10 dB
0	1	1	–20 dB
0	1	0	–30 dB
0	0	1	*
0	0	0	*

* smaller than –90 dB at 1 kHz and 10 V r.m.s. input.

Design: T. Giesberts

[944005]

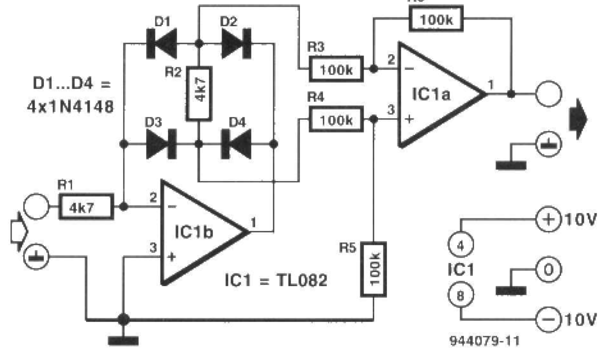


NEAR-IDEAL RECTIFIER

Occasionally, an ideal full-wave rectifier is required, for example in a dynamic compressor where switching thresholds and non-linear behaviour are not acceptable. The present circuit approaches this ideal.

The circuit consists essentially of two parts: the rectifier proper, IC_{1b}, and a differential amplifier, IC_{1a}.

The rectified voltage appears acrossR_2. Opamp IC_{1b} ensures that rectifiers D_1–D_4



function as ideal components without switching thresholds.

Since most circuits require a direct voltage with respect to earth, IC_{1a} functions as a differential amplifier. Its output carries the rectified voltage across R_2. Assuming that the supply voltage is ±10 V, the level of the alternating signal at the input must not exceed 16 V_{pp}.

The circuit draws a current of only a few milliamperes.

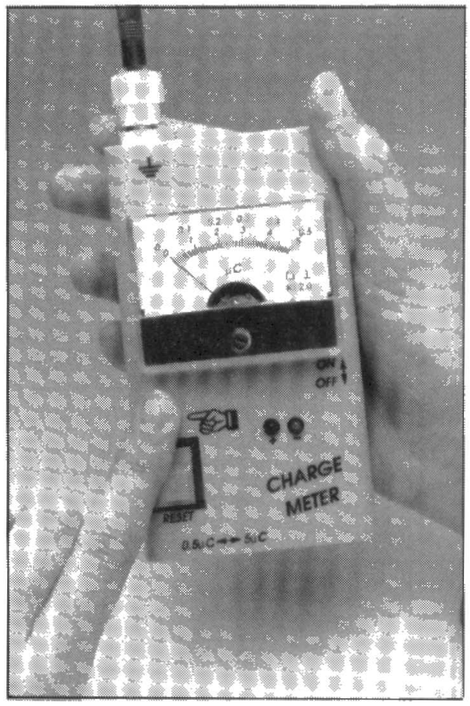
Design: H. Bonekamp

[944079]

CHARGE METER

Design by H. Bonekamp

Electrostatic discharge in a workshop or laboratory can lead to a great deal of damage. Even the electrostatic charge on the human body can give rise to voltages of up to 20 kV. Most semiconductors and integrated circuits can not withstand such potentials. With the meter described in this article the presence of unwanted electrostatic charges can be detected so that protective measures can be taken.



Although work on circuits containing semiconductors and integrated circuits should always be carried out with an earthed soldering iron and an earthing wrist strap, it is still good practice to ascertain whether anti-ESD (electrostatic discharge) measures are necessary. The meter described, an accurate electrometer with two ranges, 0–5 µC and 0–500 nC, is perfect for this task.

Design considerations

With reference to **Fig. 1**, consider an object (or person), C_1, with a certain electrostatic charge, which is used to charge a (previously uncharged) capacitor C_2. *R* is the inevitable transfer resistance. Provided that the value of C_2 is not less than 10 times that of C_1, the larger part of the charge on C_1 will have moved to C_2 within a relatively short time. The potential across C_2 can then be measured to form a measure of the charge the capacitor has received.

Since the charge is the product of capacitance and potential, $Q_1 = C_1U_1$ and $Q_2 = C_2U_2$. When the switch is closed, after a theoretically infinite time, $U_1 = U_2$, that is, U_2 may be taken as the original potential across C_1, multiplied by the ratio $C_1{:}(C_1+C_2)$. Using this in the formula for the charge on C_2:

$$Q_2 = U_1(C_1C_2)/(C_1+C_2).$$

Assuming that C_1 is negligibly small compared with C_2:

$$Q_2 = C_1U_1.$$

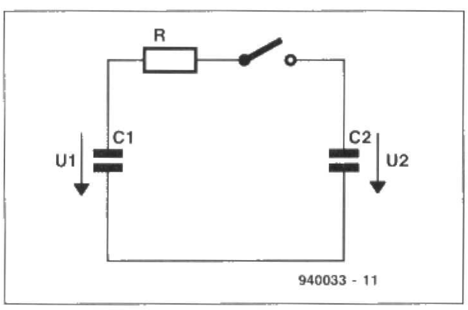


Fig. 1. The principle of the charge meter.

Resistance *R* merely delays the charge transfer, but has no effect on the amount of charge ultimately transferred.

Circuit description

In the circuit diagram in **Fig. 2**, capacitor C_1, corresponding (confusingly) to capacitor C_2 in **Fig. 1**, functions as the measuring element. Each measurement starts with closing S_1, so that C_1 is discharged very rapidly via R_1. If the (charged) object is connected to the input pin (or a person touches this pin), a transfer of charge takes place. As described earlier, C_1 will then be charged to a potential that forms a direct measure of the charge originally on the object or person. Since the 'earthy' side of C_1 is at half the supply voltage, the capacitor may be charged in a positive as well as in a negative sense. Note that the capacitance of most people is in the range of 100–200 pF, so that the value of C_1 more than meets the requirement mentioned earlier that it should be at least 10 times that of the object or person.

The potential across C_1 is applied to the non-inverting input of IC_{1a}, which, because of its high impedance, hardly constitutes a load on the capacitor. The

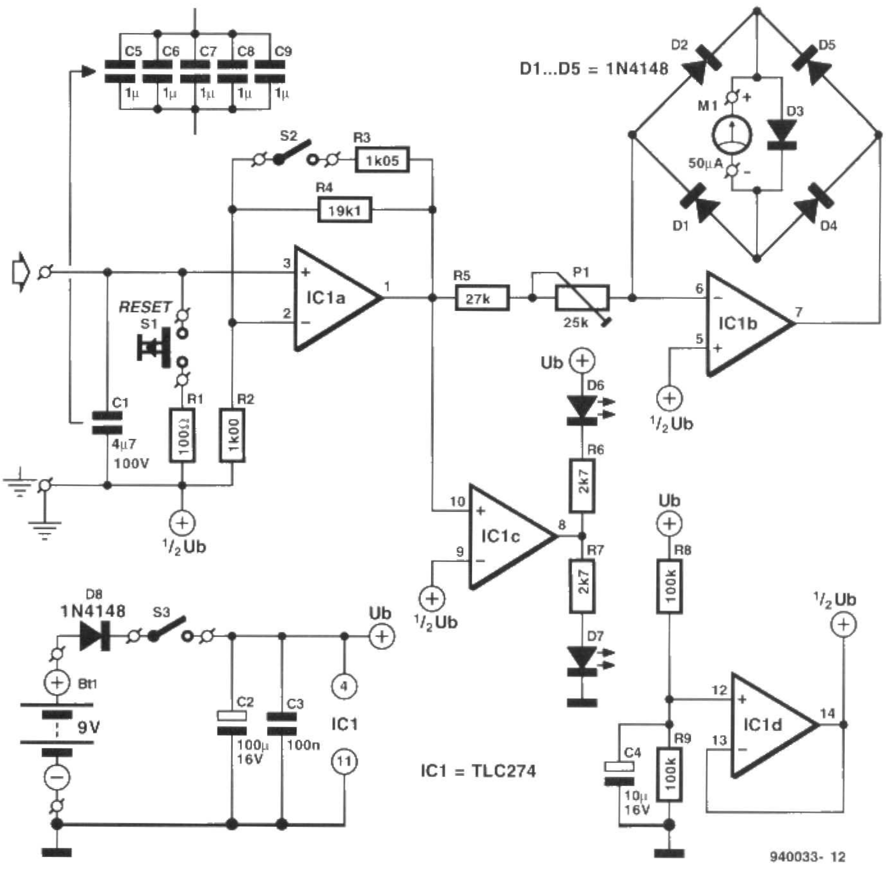


Fig. 1. Circuit diagram of the charge meter.

Electrostatic discharge (ESD)

Electrostatic discharge (ESD) may not only interfere with the correct operation of equipment, but can also lead to irreparable damage to certain components.

The discharge is obviously preceded by the forming of an electrostatic charge, which is normally caused when a conductive and a non-conductive material are rubbed together. Typical examples are the human body and nylon carpets or car upholstery, terylene clothing and plastic wrapping material. The consequent electrostatic voltage may be as high as 20 kV.

The phenomenon depends strongly on the humidity of the ambient atmosphere. The drier the air, the higher the voltage. Research has shown that in this respect the human body may be considered as a capcitor with a value of 100–200 pF. During an ESD, people have an internal resistance of 150–1500 Ω. The electrostatic charge appears to be concentrated in protruding parts of the body, such as the hands. Because of this and the speed of the discharge pulses, ESD can rapidly damage semiconductors when these are touched by hand.

Recognition of the phenomenon is not easy, because shocks are felt only when the potential is higher than 2–3 kV, but this level of voltage can already seriously damage sensitive semiconductors.

Electrostatic charges may be prevented by replacing non-conducting carpets, clothing or packing by conducting materials. Also, the use of an earthed wrist strap is advisable when semiconductors are handled.

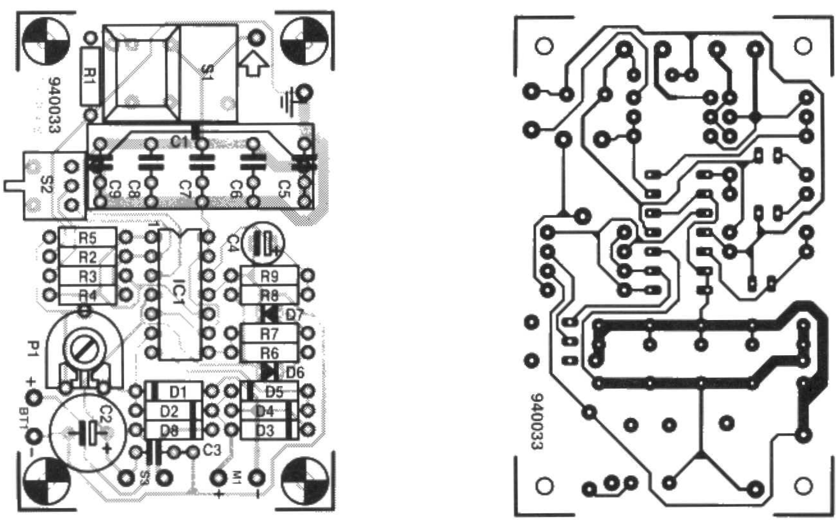


Fig. 3. Printed circuit board for the charge meter.

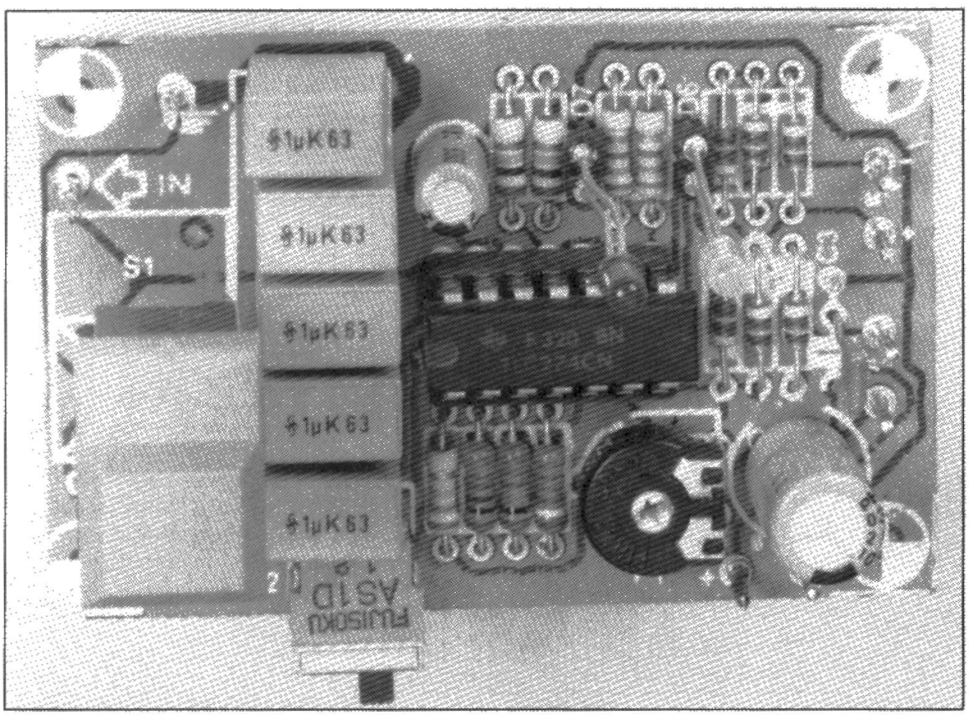


Fig. 4. Completed printed circuit board.

amplification of IC_{1a} can be switched between ×2 and ×20 by S_2, which switches R_3 into the feedback loop, or disconnects it from the loop.

From the output of IC_{1a}, the signal takes two paths: one to the metering network and the other to comparator IC_{1c}.

The inverting input of IC_{1c} carries half the supply voltage as reference potential. In this way, the output of the comparator indicates the polarity of the measured charge. If this is negative, D_6 lights and when it is positive, D_7 lights.

The part of the output of IC_{1a} fed to the metering network is applied via R_5 and P_1 to IC_{1b}, which drives the meter, M_1. The diodes ensure that the meter pointer always deflects in the same direction, irrespective of the polarity of the original charge. Diode D_3 protects the meter against too high a potential across it.

Power is derived from a 9 V battery, which is buffered by C_2 and decoupled for r.f. by C_3. S_3 is the on-off switch, while D_8 protects the circuit against a wrongly connected battery.

The stabilized half supply voltage mentioned on a few earlier occasions is obtained from potential divider R_8-R_9 and buffer stage IC_{1d}. It is essential for the correct working of the circuit that the output of IC_{1d} is connected to a good earthing point (mains earth or water pipe).

Construction

The charge meter is best built on the printed-circuit board shown in **Fig. 3**. The design of the board allows either a single 4.7 μF capacitor (C_1) or five 1 μF capacitors (C_5–C_9) in parallel to be used as the input capacitor (see **Fig. 4**).

When the board has been completed, mount it in a small case as shown at the beginning of this article and in **Fig. 5**. A suggested front panel is shown in **Fig. 6** (which is not available ready made).

It is absolutely essential that the circuit is properly earthed, either to the mains earth or to a water pipe.

The input (touch) terminal can be, for instance, a non-insulated audio socket.

Calibration and use

A variable power supply is required for calibrating the meter.

- Set the meter to the 5 μC range (S_2 closed).
- Adjust the variable power supply to give an output of exactly 1.06 V if C_1 is used and of 1.0 V if C_5–C_9 are used. Check the output with a digital voltmeter.
- Connect the output of the variable power supply across C_1 or C_5–C_9, as the case may be.
- Adjust P_1 for fulll-scale deflection (f.s.d.) on M_1.

When the earth wire has been connected, the power supply has been switched on,

and reset knob S_1 has been pressed, the charge meter is ready for use. The meter should then be at 0. Touch the input, or connect an object suspected of having a static electric charge to the input.

Never forget to press the reset button before making a measurement. Also, it is advisable to start a measurement with the meter set to the higher range.

Parts list

Resistors:
R_1 = 100 Ω
R_2 = 1.00 kΩ, 1%
R_3 = 1.05 kΩ, 1%
R_4 = 19.1 kΩ, 1%
R_5 = 27 kΩ
R_6, R_7 = 2.7 kΩ
R_8, R_9 = 100 kΩ
P_1 = 25 kΩ preset potmeter

Capacitors:
C_1* = 4.7 μF, 100 V, polystyrene
C_2 = 100 μF, 16 V, radial
C_3 = 100 nF
C_4 = 10 μF, 16 V
C_5–C_9* = 1 μF, 63 V, polystyrene

* alternatives – see text

Semiconductors:
D_1–D_5, D_8 = 1N4148
D_6 = LED, green, low current
D_7 = LED, red, low current

Integrated circuits:
IC_1 = TLC274CN

Miscellaneous:
S_1 = miniature spring-loaded push-button switch
S_2 = slide switch, 1 make contact , PCB model
S_3 = single-pole, single-throw switch
M_1 = moving-coil meter, 50 μA
Bt_1 = 9 V battery with terminal clip
PCB Ref. 940033 (p. 110)

[940033]

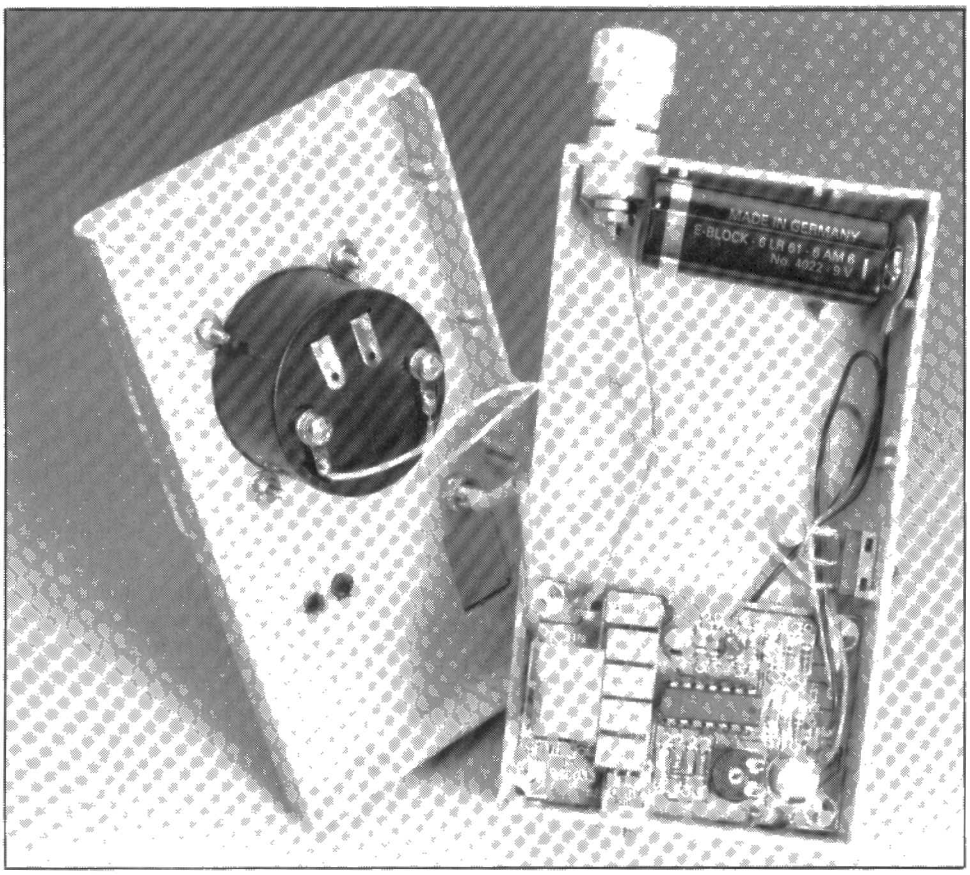


Fig. 5. Suggested final assembly of the charge meter.

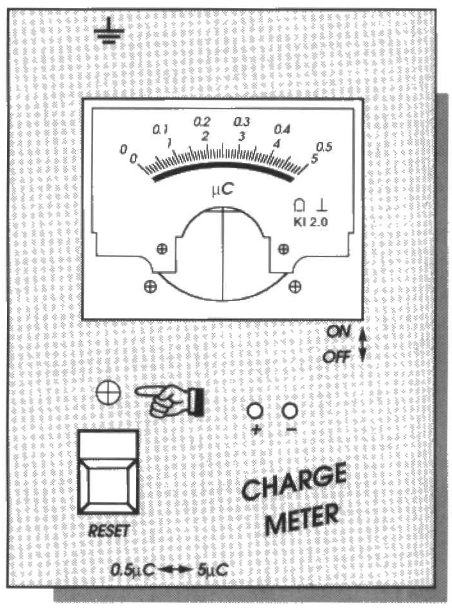


Fig. 6. Suggested front panel.

APPLICATION NOTE

The content of this note is based on information received from manufacturers in the electrical and electronics industries, or their representatives, and does not imply practical experience by *Elektor Electronics* or its consultants.

SOLID-STATE TEMPERATURE SENSOR TC62X

By G. Kleine

Temperature sensors and switches in the TC62x series from Teledyne Components make some interesting applications practicable. They are particularly suited for applications where, owing to shock and vibration, mechanical devices can not be used. Both sensors and switches operate from supply voltages in the range 4.5–18 V.

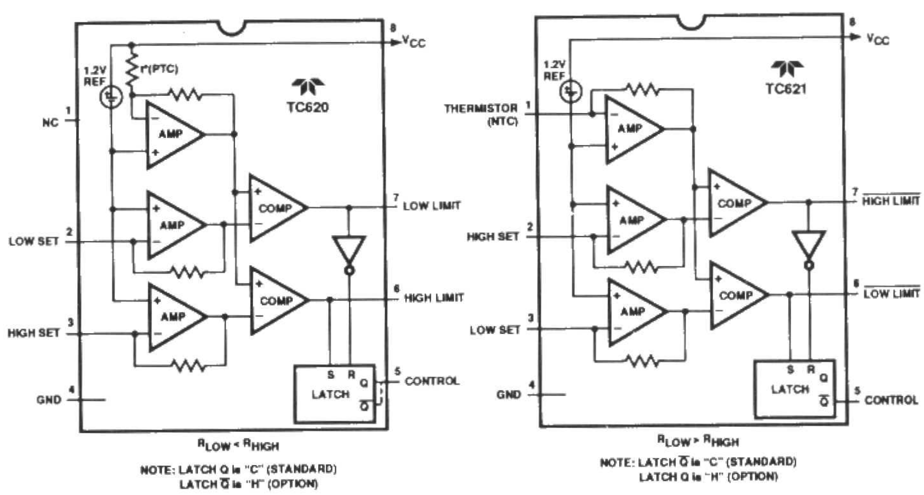
The TC620/621 can react to an upper and a lower temperature setting with the aid of two resistors. Three outputs serve to control the external load (fan, heating).

The TC626 is a 3-pin thermal switch intended for operation in the frequency range 0–125 °C.

The accuracy of the switching temperatures, according to the manufacturers's data, is ±3 °C.

FUNCTIONAL DIAGRAM

1



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Sensors TC620/621

The internal circuit of sensors TC620 and TC621 is shown in **Fig. 1**. Whereas in the TC620 an internal PTC (positive temperature coefficient) detector converts changes in the temperature into changes in resistance, the TC621 operates with an external NTC (negative temperature coefficient) resistor as detector. This makes the operation independent of the chip temperature, and makes it possiblel for the detector to be sited in the most convenient position for the particular application.

Inputs LOW SET (pin 2) and HIGH SET (pin 3) serve to set the two threshold temperatures. This only requires a resistor to the +ve supply line, V_{cc}. The value of the resistor determines the position of the threshold according to the formulas:

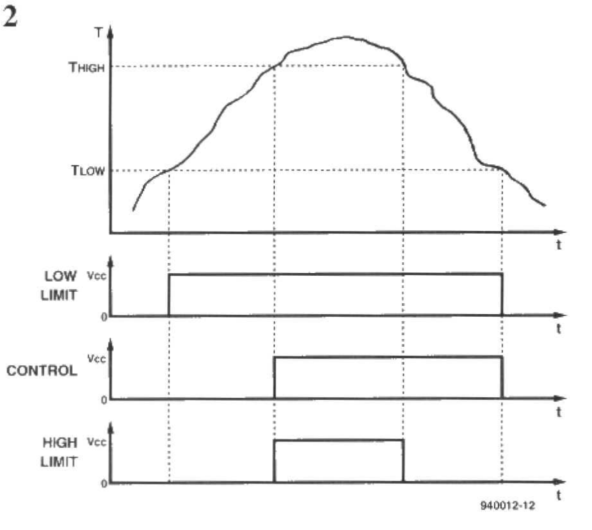
$$R = 0.783T + 91 \qquad (T < 70\ °C)$$

$$R = T + 77 \qquad (T > 70\ °C)$$

where R is the resistance in kΩ and T is the threshold temperature in °C.

The TC620/621 circuits have three outputs: LOW LIMIT, HIGH LIMIT, and CONTROL. The first two are used to indicate whether the relevant threshold is being exceeded. In the TC620 they are logic low when the ambient temperature is below the threshold and logic 1 when the ambient temperature is above the threshold. The third output, in conjunction with a bistable, has an hysteresis function.

2



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3

T/°C	R/kΩ
0	91
10	99
20	107
30	115
40	122
50	130
60	138

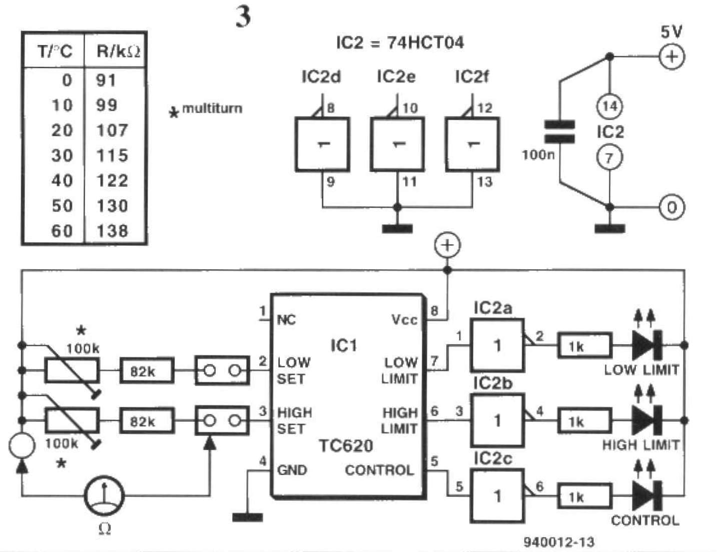

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Table 1

Type	Case	Detector	Control
TC620	Cxxx	internal	normal polarity
TC620	Hxxx	internal	inverse polarity
TC621	Cxxx	external	normal polarity
TX621	Hxxx	external	inverse polarity

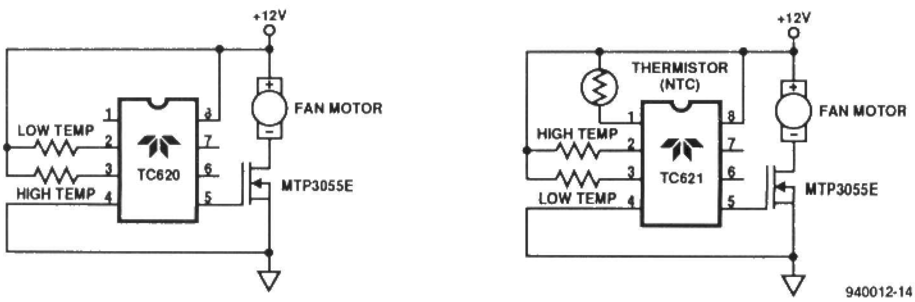
xxx =	COA =	8-pin SOIC	0 °C – 70 °C
	EOA =	8-pin SOIC	–40 °C – +85 °C
	CPA =	8-pin plastic DIP	0 °C – 70 °C
	EPA =	8-pin plastic DIP	–40 °C – +85 °C
	MJA =	8-pin ceramic DIP	–55 °C – +125 °C

Table 2

Type	Case	Ambient temperature
TC626 yyy CAB	3-pin TO-220	0 °C – 70 °C
TC626 yyy EAB	3-pin TO-220	–40 °C – +85 °X
TC626 yyy VAB	3-pin TP-220	–40 °C – +125 °C
TC626 yyy CZB	3-pin TO-92	0 °C – +70 °C
TC626 yyy EZB	3-pin TO-92	–40 °C – +85 °C
TC626 yyy VZB	3-pin TO-92	–40 °C – +125 °C

where yyy indicates the switching temperature.
Switching temperatures between 0 °C and 125 °C are available in 5 °C steps; for instance, a Type TC626 045 VAB is housed in a TO-220 case.and has a switching threshold of +45 °C.

4



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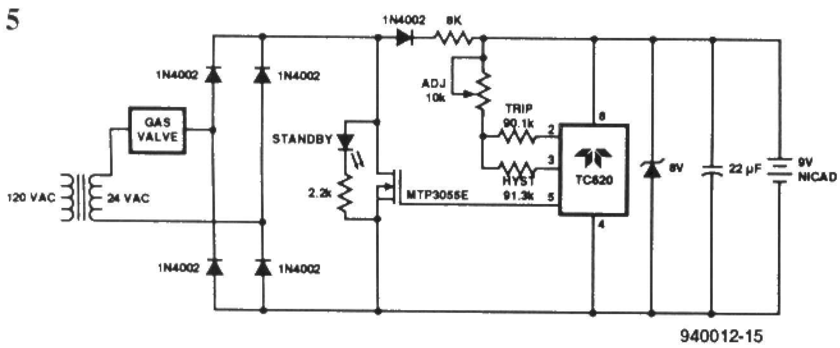

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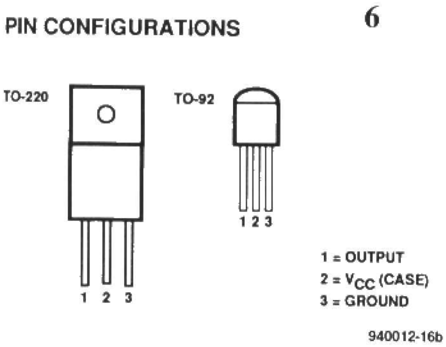
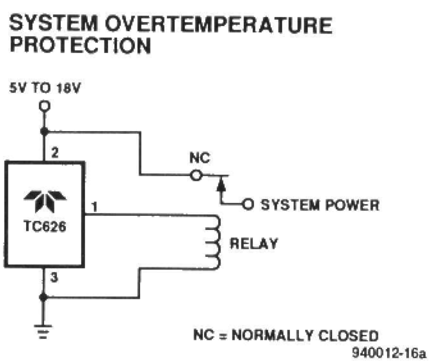




When the HIGH LIMIT threshold temperature is exceeded, the bistable is set, whereupon it can start, say, a cooling system via pin 5 (that is, the CONTROL output), which is then logic high. Only after the temperature has dropped below the LOW LIMIT threshold again, is the bistable reset (when the CONTROL output becomes logic low)—see **Fig. 2**.

In the Type TC620-H the CONTROL output is active low.

The operation of the TC621, owing to its external NTC resistive detector, is exactly the opposite. Compared with the TC620, its HIGH LIMIT and LOW LIMIT are reversed, as are the allocation and polarity of the three outputs. This means that in this IC the output goes logic low when the ambient temperature exceeds the set threshold. The CONTROL output will have the same polarity as that in the TC620 if the other Q output of the bistable is used. Thus, pin 5 becomes logic high when the HIGH SET threshold is exceeded and logic low when the temperature is below the LOW SET threshold.

Table 1 gives a correlation of the various versions of the TC620 and TC621 and the relevant temperature ranges.

The TC620 may be tested with the simple circuit shown in **Fig. 3**. The two resistors that determine the threshold temperatures are connected in series with preset potentiometers, which enable accurate settings to be obtained (set the resistance with the aid of an ohmmeter).

Because of the PTC detector, it is important that the chip does not warm up. Therefore, the manufacturers permit an output current of only 1 mA. In the prototype circuit, each output is connected to a driver stage contained in IC_2.

Since the TC621x uses an external NTC detector, its output current can be up to 10 mA.

Figure 4 shows how simple an application circuit using a TC620/621 chip can be. Here, dependent on the ambient temperature, a fan is switched on and off. The CONTROL output of a TC620 drives a power MOSFET, which functions as the on/off switch for the fan.

Figure 5 shows how a thermostat is switched by a TC620. In the diagram, the power MOSFET switches a valve in the heating system. The circuit is powered by the 24 V supply of the heating system. When the power MOSFET is on, a standby battery comes into circuit.

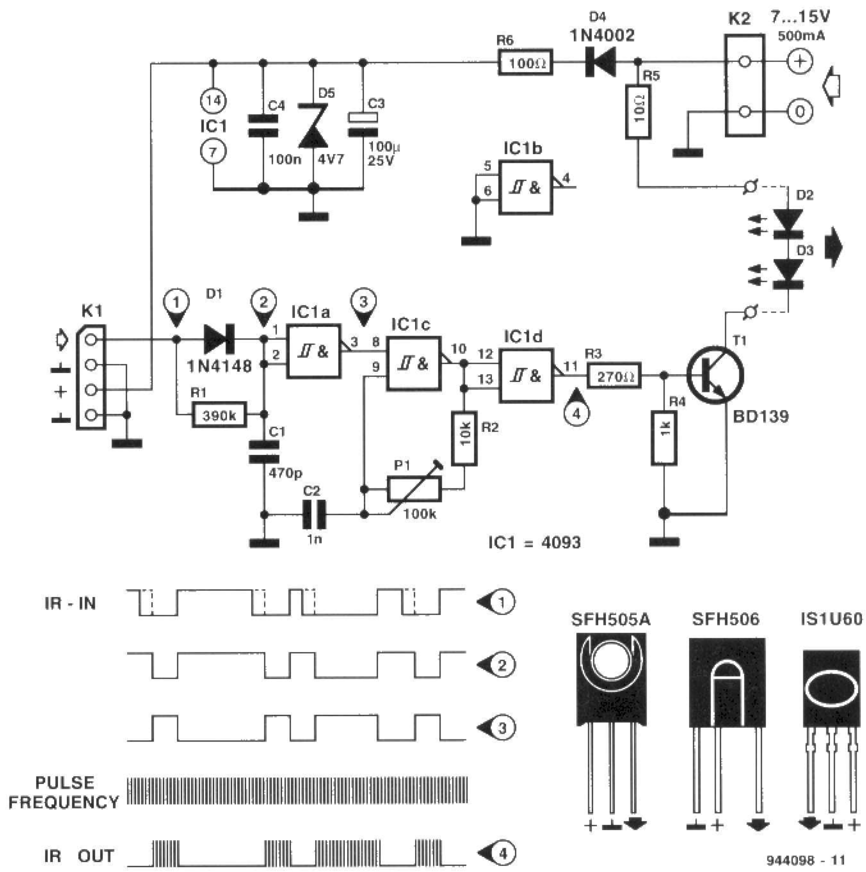
Thermal switch TC626

The Type TC626 thermal switch is available for operation in the temperature range 0–125 °C. It is housed in a TO-220 or TO-92 case—see **Fig. 6a**. The switching output delivers op to 10 mA in the TO-92 version, and up to 50 mA in the TO-220 version. **Table 2** gives an overview of the various versions and the associated temperature ranges.

The switching output of the TC626 is logic low as long as the threshold temperature is not exceeded. When the ambient temperature reaches the threshold, the output becomes logic high. This makes it possible, for instance, to protect an apparatus by having a relay switch off its supply voltage when a given temperature is exceeded—see **Fig. 6b**.

[940012]

GENERAL PURPOSE INFRA-RED BUFFER



The buffer enables 'stretching' the signal of any infra-red (IR) remote control. This makes it possible, for instance, to control a video recorder in the living room from the bedroom or kitchen. The high-frequency modulated binary signal is converted by an IR receiver/decoder (see point 1 in the diagram) into a serial signal without that high frequency

The trailing edge of the received signal is delayed by R_1-C_1, since the IR receiver delays the leading edge. Without this compensation, the bits would become slightly wider than they were originally. This may result in errors in the receiver which ultimately must decode the signals, or a contraction of the distance to be spanned. Owing to D_1, the *RC* network does not affect the leading edge (2).

To retransmit the decoded signals as IR signals, the lower periods of the signal must be filled with the basic frequency of the detected signal, which in the RC-5 code is 36 kHz. This is done with the aid of IC_{1a}, IC_{1c} and IC_{1d}.

When pin 8 of IC_{1c} is high, this gate will generate the carrier frequency with the aid of R_2, P_1 and C_2. This signal is used via gate IC_{1d} to switch output transistor T_1. This high frequency is transferred when the received signal is low, because the output of IC_{1a} is then high.

Gate IC_{1d} is needed to switch the transistor off when oscillator IC_{1c} is disabled.

Fit D_2 and D_3 close to the equipment that is to receive the IR signals. The link between the LEDs and the present circuit can be simply of loudspeaker cable.

The power supply may be a mains adaptor that can deliver a current of about 0.5 A.

Preset P_1 may be adjusted by 'ear': simply vary it until the largest required distance can be spanned. If an oscilloscope is available, compare the frequency of the original signal with that of the oscillator and adjust P_1 until they are identical.

Design: A. Rietjens
[944098]

PROGRAMMABLE PULSE SPACING METER

The meter can measure the spacing between the trailing edges of not fewer than two, nor more than, nine pulses in a train. The number of pulses is selected by a rotary switch.

The interval is measured with the aid of a 1 MHz pulse generator, IC_4. During measurements the generator signal is present also at the output of the circuit to enable an external counter to be used.

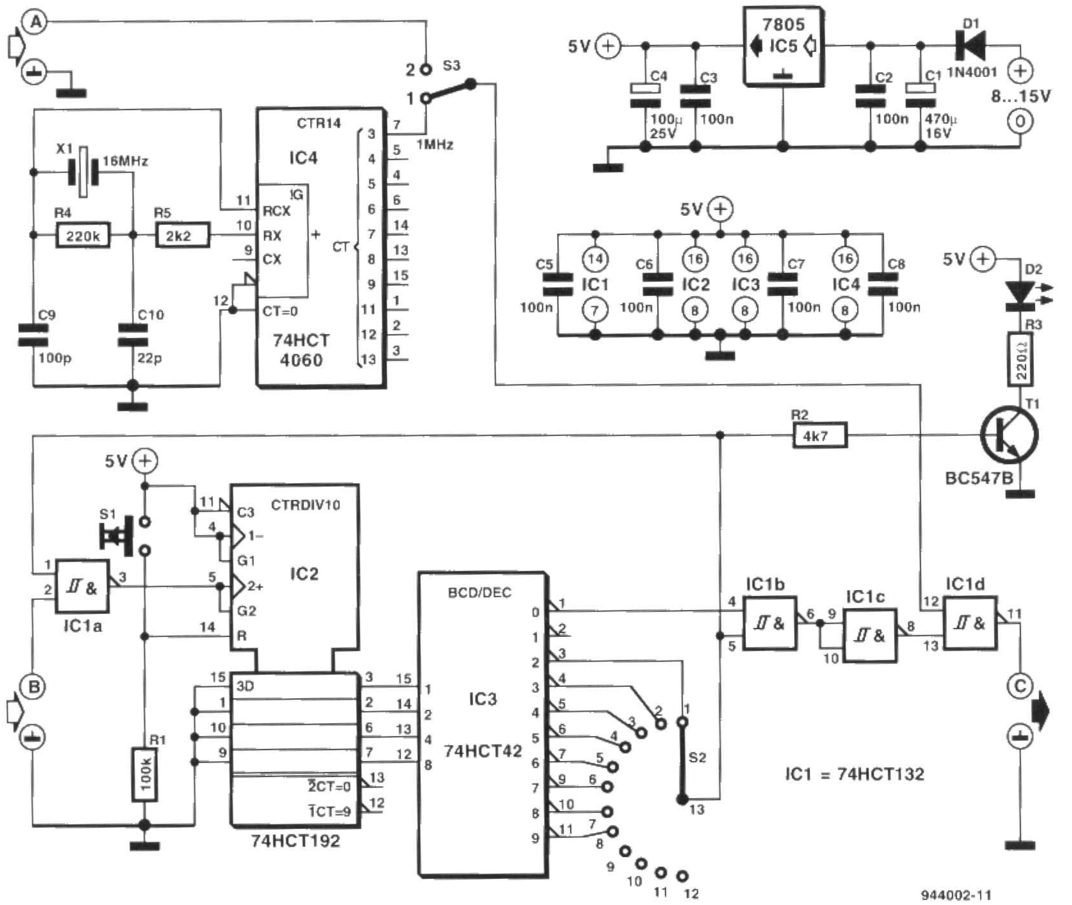
Decade scaler IC_2 is reset with switch S_1, whereupon pin 1 (output 0) of BCD-to-decimal converter IC_3 goes, while the level at all the other outputs of this circuit go high. This results in the output of IC_{1d} becoming high, so that D_2 lights. The 1 MHz pulses from IC_4 (arranged as a :16 divider), or the external generator signal, depending on the position of S_3, are then blocked by IC_{1d}.

On the trailing edge of the first pulse that reaches the circuit after this has been reset, IC_2 is set to position 1. The level at pin 1 of IC_3 then goes high, resulting in the clock signal appearing at the output of the circuit. On the trailing edge of the second pulse, pin 3 of IC_3 goes low, whereupon IC_{1d} blocks the clock pulses if rotary switch S_2 is in the position shown in the diagram. Setting this switch to a different position determines at which clock pulse the transfer of the 1 MHz pulse train is blocked. The pulse input is then disabled via the pole of S_2 and gate IC_{1a}.

The counter connected to the output of the circuit indicates how much time has elapsed between the first and the next desired (2–9) pulse. At the specified reference frequency this will be in steps of 1 μs. If the clock signal were 1 kHz, the steps would be 1 ms.

The circuit draws a current of about 20 mA, which makes battery operation possible.

Design: K. Dietrich
[944002]



DRIVE FOR BISTABLE RELAY

Bistable relays have the great advantage that once they are in the wanted position they do not need power to maintain that status. Another advantage is that their position is retained when the power fails or is switched off. This means that such a relay can be used as a semi-permanent memory.

The present circuit is intended to drive a bistable relay. Typical of such relays is that they have two windings.

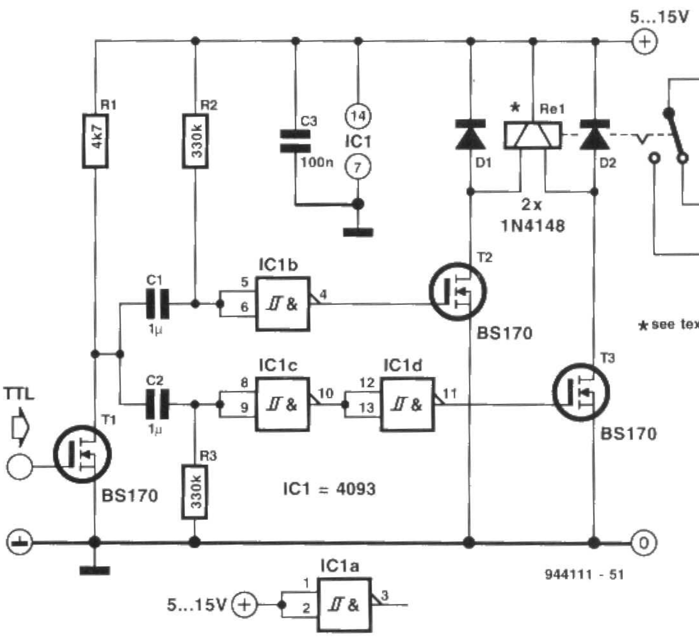
Since only a short voltage pulse is needed to switch the relay to its wanted position, the drive circuit contains two monostable multivibrators (MMV). One of these consists of IC_{1b}, C_1 and R_2; the other of gates IC_{1c} and IC_{1d} and C_2 and R_3. Buffers T_2 and T_3 ensure that the MMVs can provide sufficient current.

The digital input signal is buffered and inverted by T_1, so that a leading edge of the signal at the base rises, whereas that of the signal at the drain decays. A decaying leading edge triggers MMV IC_{1b}, whereupon the output of IC_{1b} briefly goes high. This level switches on T_2, so that the associated winding of Re_1 is energized and the relay switches. A trailing edge at the input causes the other MMV to be triggered, which results in the other winding of the relay being energized. The relay then switches to its second position.

Diodes D_1 and D_2 protect the output transistors agains voltage peaks caused by the switching of the relay.

Design: G. Kleine

[944111]



SERIAL 12-BIT A-D CONVERTER

The MAX187 is eminently suitable for building a good-quality 12-bit A-D converter. Communication with the computer is serial, but not via the standard RS232 protocol. All that is needed are three free I/O lines. In the prototype, the Centronics port was used.

The diagram shows a thermometer designed around the MAX187 with an AD592 serving as the sensor. Because of the potential across IC_2 and resistor R_1, a supply voltage of 8 V is required. The A-D converter works from 5 V, however. This is why there are two voltage regulators. For other applications, IC_1 can, therefore, be omitted. Although the circuit draws a current of only a few mA, the power supply is not taken from the Centronics port.

The A-D converter is enabled by making pin 3 high. If this pin is left open, the IC is enabled, but the internal reference voltage (4.096 V) is disabled. In that case, an external reference voltage must be applied to pin 4.

A start-of-conversion pulse is generated by making pin 7 low. This pin must remain low until the conversion data have been read. During the conversion, the clock input (pin 8) must be low. The data output (pin 6) is high impedance when pin 7 is high. Pin 6 is low as long as the conversion lasts, but goes high as soon as it is ended.

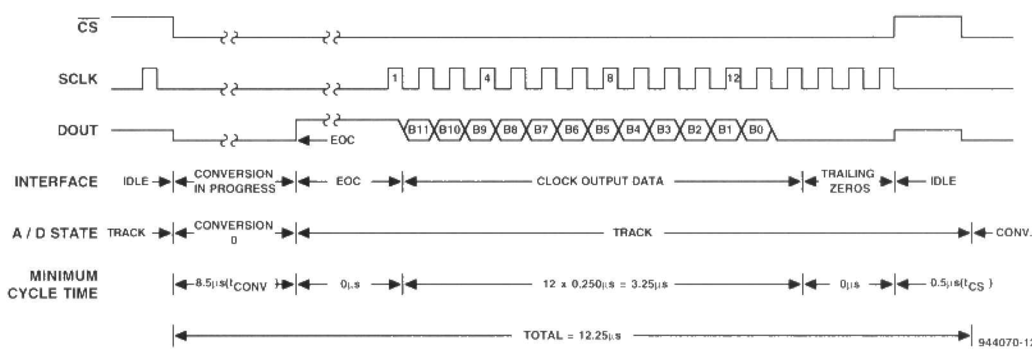
The associated software must be able to detect the going high of pin 6, whereupon it must commence with reading the 12 data bits, starting with the MSB. Thirteen clock pulses at pin 8 are needed for this. The data change at the trailing edges of the clock.

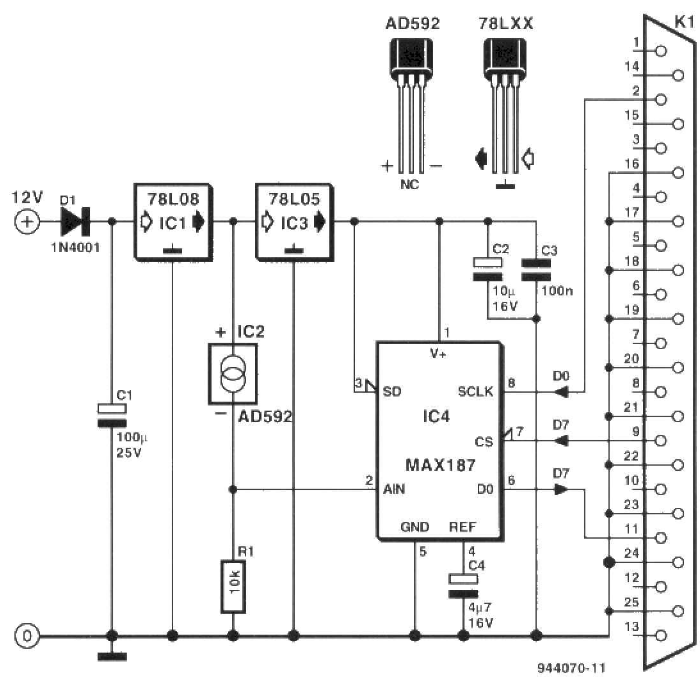
The MAX187 is quite fast: 8.5 µs for the conversion; 13 times 0.25 µs for reading and a pause of 0.5 µs: a total conversion time of 12.25 µs. See also the timing diagram.

For use as a thermometer, the QBASIC program shown indicates the measured temperature on the monitor. QBASIC is delivered free of charge with MOS-DOS versions 5 and 6.

```
maxwrite = &H378
maxread = maxwrite+1

WHILE 1
  OUT maxwrite, 128: REM CS high
  OUT maxwrite, 0: REM start conversion

  WHILE INP(maxread) AND 128 = 0: REM wait for EOC (input inverted!)
  WEND

  OUT maxwrite, 1: REM clock high, extra clock to start
  OUT maxwrite, 0: REM clock low
  total = 0
  FOR clocks = 11 TO 0 STEP -1
    OUT maxwrite, 1
    value% = (NOT INP(maxread)) AND 128: REM inverted input!
    IF value% <> 0 THEN total = (2 ^ clocks) + total
    OUT maxwrite, 0: REM next clock
  NEXT
  PRINT total

WEND
```

944070-13





Design: K.M. Walraven

[944070]

DISCRETE PREAMPLIFIER

Quality-conscious audio buffs still prefer discrete designs. And quite rightly so, because although there are very good operational amplifiers available, discrete designs offer just that little bit extra.

The present preamplifier is a symmetrical Class A design. The input is a double differential amplifier consisting of dual transistors Type MAT02 or MAT03. A stable d.c. operating point is ensured by current sources T_3 and T_4, which use LEDs as reference—D_1 and D_2 respectively.

The current through the LEDs is held stable by current source T_5. It is essential for good thermal stability that the transistors and associated diodes (T_3 and D_1, and T_4 and D_2) are mounted in close contact.

The input signals are applied to push-pull drivers T_6 and T_7, which feed the output stages, consisting of emitter followers T_{10} and T_{11}. Transistors T_8 and T_9 ensure a constant quiescent current through the emitter followers. It is necessary for good thermal stability that T_8 and T_{10}, and T_9 and T_{11}, are mounted in close contact. To this end, their flat sides, with heat conducting paste in between, are juxtaposed. The pairs are held together with a loop of bare copper wire.

Before the mains is switched on, set P_1 to maximum resistance. Switch on the mains, wait for about a minute and then adjust P_1 for a quiescent current through T_{10} and T_{11} of 15 mA, corresponding to a voltage drop of 150 mV across R_{23} and R_{24}.

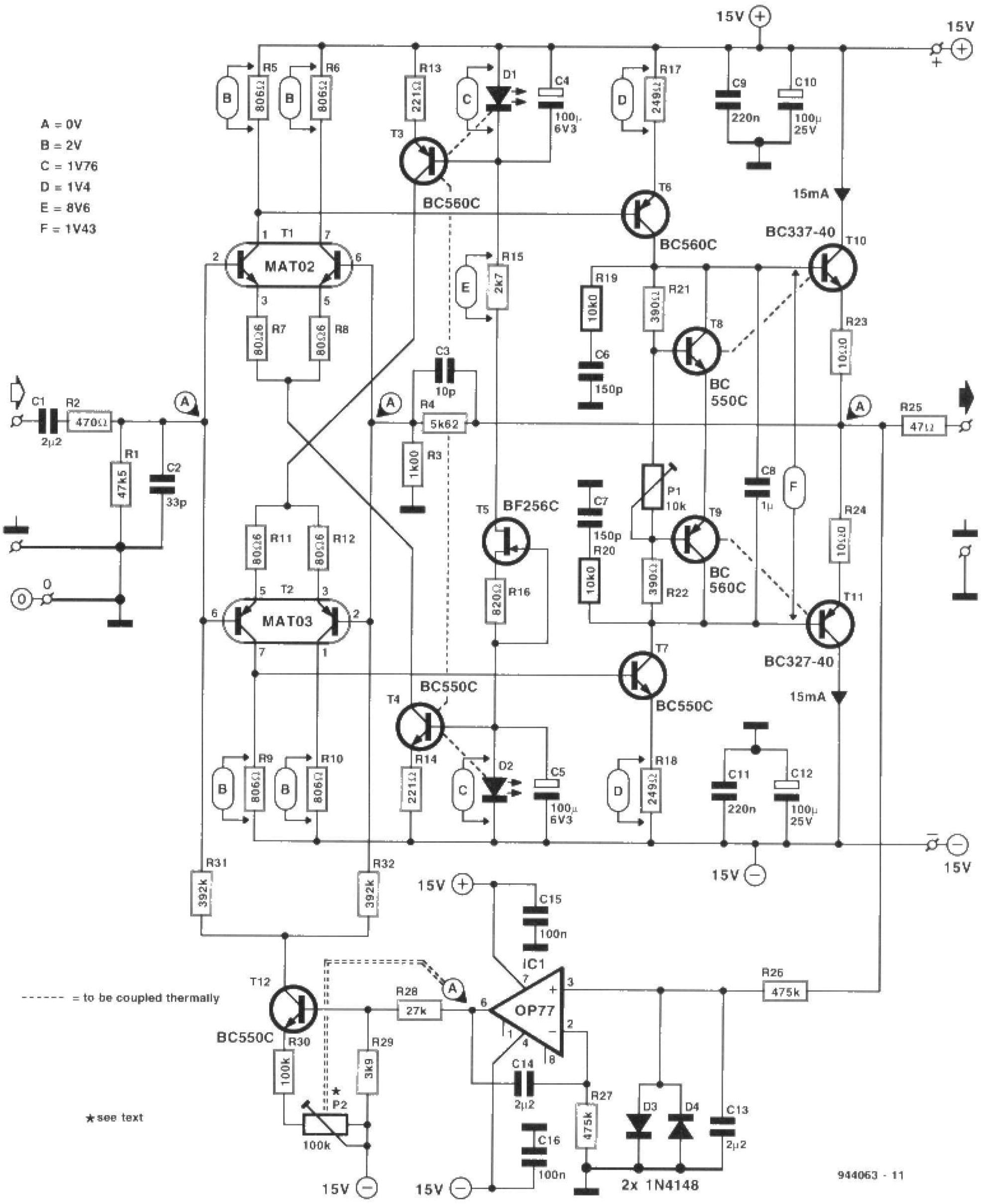


Fig. 1. Circuit diagram of the discrete preamplifier

Since the amplifier is d.c. coupled throughout, the likelihood of a fairly high direct voltage at the output would be great, the more so because the input transistors are not truly complementary. This is, however, obviated by an active d.c. correction that holds the direct voltage at the output at zero in all circumstances. For this purpose, the output signal is passed via low-pass filter R_{26}-C_{13} to integrator IC_1. Tis arrangement does not affect fast variations of the signal. If, however, the output signal has a d.c. component, T_{12} will conduct to some degree, so that the bases of T_1 and T_2 are pulled into a negative direction. In a negative direction, because T_1 (n-p-n) has an inherently greater voltage amplification (×3) than T_2 (p-n-p).

Adjust P_2 immediately on switch-on for as low a direct voltage at the output as possible. From then on, any variations caused by temperature changes will be corrected by IC_1. The speed at which the correction takes place can be increased by giving R_{26} and R_{27} lower values.

It is important for optimum

Some parameters

(measured with an output of 1 V r.m.s. across 47 kΩ)

THD	≤0.00005% (at 1 kHz)
	≤0.0004% (at 20 kHz)
THD + N (B = 22 Hz–80 kHz)	<0.0012% (20 Hz–20 kHz)
Signal-to-noise ratio (B = 22 Hz–22 kHz)	>104 dB
Bandwidth	1.5 Hz–3.7 MHz
Slew rate	about 200 V μs^{-1}
Rise time	about 0.1 μs
Input impedance	47 kΩ
Sensitivity	150 mV
Peak output voltage	about 9 V r.m.s.

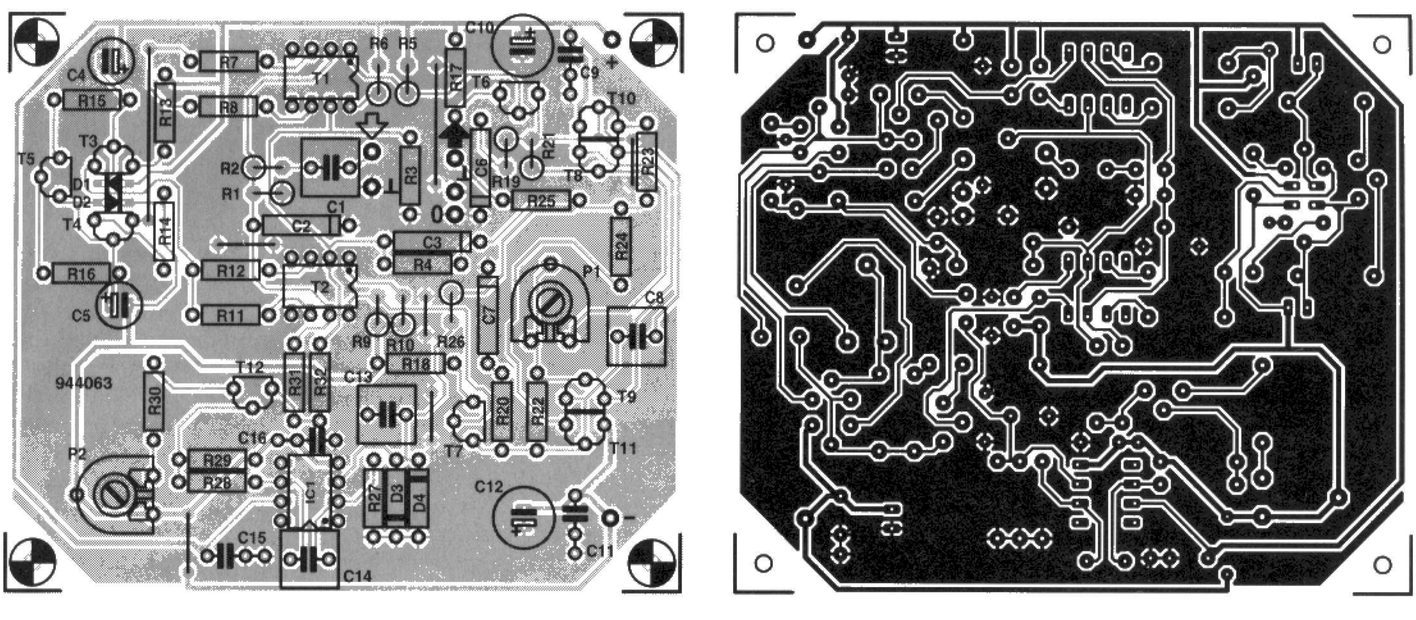


Fig. 2. Printed circuit board for the discrete preamplifier.

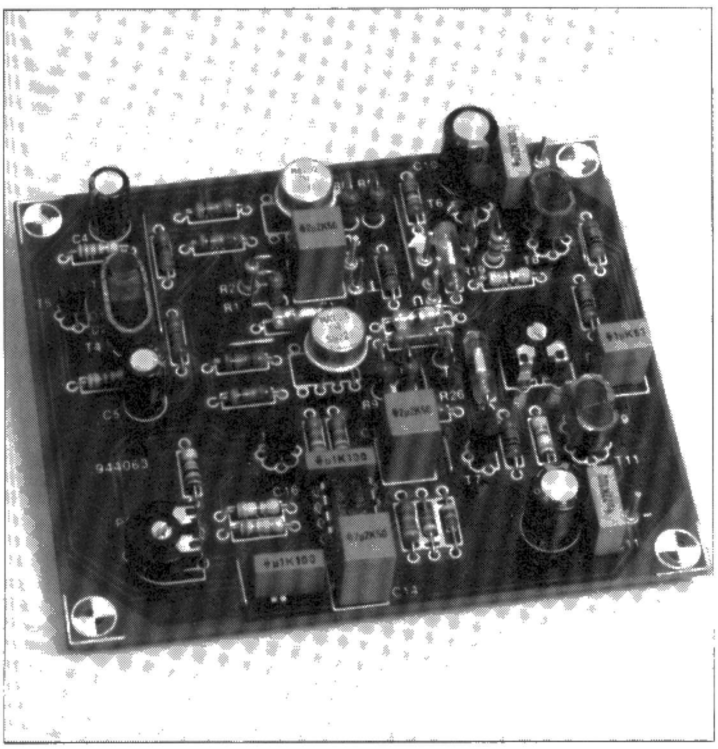


Fig. 3. Completed printed circuit board.

symmetry that the currents through T_1 and T_2 (and thus the voltage drops across R_9 and R_{10}) are equal. This can only be if the potentials across D_1 and D_2 are equal, and it is, therefore, advisable to match these diodes for equal voltage with a test current through them of 3 mA. When the diodes are matched, the drops across R_{13} and R_{14} should not differ by more than a few millivolts.

The same applies to T_6 and T_7: for good symmetry they should be matched for equal base/emitter voltage, with a current through them of 5 mA. This matching can not be done in the circuit, because the voltage drops across R_{17} and R_{18} will be equal whatever, otherwise the output would not be zero.

Low-pass filter R_2-C_2 is desdigned for maximum slew rate at a cut-off point of 9–10 MHz. If this large bandwidth results in high sensitivity to interference, it may be advisable to lower the cut-off point. If the value of C_2 is increased to 680 pF, the cut-off point drops to about 400 kHz. At the same time, the slew rate deteriorates to about 20 V μs^{-1}.

The preamplifier is best built on the PCB in **Fig. 2**, which is available ready-made.

The supply lines should be stabilized by a suitable voltage regulator.

Parts list

Resistors:
R_1 = 47.5 kΩ, 1%
R_2 = 470 Ω
R_3 = 1.00 kΩ, 1%
R_4 = 5.62 kΩ, 1%
R_5, R_6, R_9, R_{10} = 806 Ω, 1%
R_7, R_8, R_{11}, R_{12} = 80.6 Ω, 1%
R_{13}, R_{14} = 221 Ω, 1%
R_{15} = 2.7 kΩ
R_{16} = 820 Ω
R_{17}, R_{18} = 249 Ω, 1%
R_{19}, R_{20} = 10.0 kΩ, 1%
R_{21}, R_{22} = 390 Ω
R_{23}, R_{24} = 10.0 Ω, 1%
R_{25} = 47 Ω
R_{26}, R_{27} = 475 kΩ, 1%
R_{28} = 27 kΩ
R_{29} = 3.9 kΩ
R_{30} = 100 kΩ
R_{31}, R_{32} = 392 kΩ, 1%
P_1 = 10 kΩ preset potmeter
P_2 = 100 kΩ preset potmeter

Capacitors:
C_1, C_{13}, C_{14} = 2.2 μF, 50 V, pitch 5 mm
C_2 = 33 pF, 160 V, polystyrene
C_3 = 10 pF, 160 V, polystyrene
C_4, C_5 = 100 μF, 6.3 V, radial
C_6, C_7 = 150 pF, 160 V, polystyrene
C_8 = 1 μF, pitch 5 mm
C_9, C_{11} = 220 nF
C_{10}, C_{12} = 100 μF, 25 V, radial
C_{15}, C_{16} = 100 nF

Semiconductors:
D_1, D_2 = LED, red, flat
D_3, D_4 = 1N4148
T_1 = MAT02
T_2 = MAT03
T_3, T_6, T_9 = BC560C
T_4, T_7, T_8, T_{12} = BC550C
T_5 = BF256C
T_{10} = BC337-40
T_{11} = BC327-40

Integrated circuits:
IC_1 = OP77

Miscellaneous:
PCB Ref. 944063 (p. 110)

Design: T. Giesberts

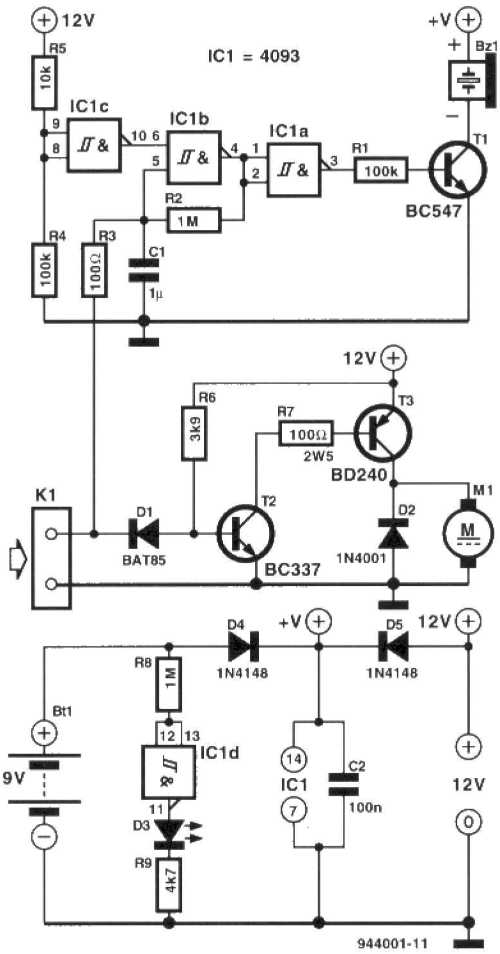
[944063]

HEAT SINK MONITOR

The monitor is intended especially for force-cooled power resistors mounted on a heat sink. Such constructions are used, for instance, to test audio output amplifiers. Since it may be forgotten to switch the associated fan on during testing, the resistors can overheat with all the consequences of this.

It is, of course, possible to use in series with the resistors a relay that is energized only when the power to the fan is switched on. This is, however, not feasible with low-value resistors (e.g.,≤1 Ω) because the transfer resistance of the relay may then play too large a role.

Another method is the use of a battery-powered alarm, which actuates a buzzer when the temperature rises above 80 °C. When the mains to the fan is switched on, the alarm is disabled. The present circuit then ensures that the fan is operated when the temperature reaches 80 °C. The advantage of this arrangement is that there is no (electrical) fan noise at low powers.



The circuit is powered by battery Bt_1 via diode D_4 if there is no mains voltage present. Stage T_3 is not powered and the logic level at the inputs of gate IC_{1c} is 0. This enables oscillator IC_{1b}, but this can not yet oscillate because C_1 is short-circuited by a thermal switch connected to K_1. The output of IC_{1a} is thus low, so that T_1 is off and the (direct-current) buzzer is not energized. When the temperature rises above 80 °C, the thermal switch opens and IC_{1b} begins to oscillate, whereupon the buzzer sounds intermittently.

When the mains is switched on, a 12-V potential is applied to the emitter of T_3, whereupon the level at the inputs of IC_{1c} goes high. This causes the oscillator to be disabled, so that the buzzer is deenergized. At temperatures below 80 °C, the sensor contacts are closed, so that T_2 and T_3 are switched off. Above that temperature, the sensor contacts open, whereupon T_2 arranges for the fan to be switched on via T_3. When the temperature drops below 80 °C, the sensor contacts close and the fan is switched off.

Gate IC_{1d} provides a low-battery indication (which operates only when the mains is switched on).

The sensor must be fitted on to the heat sink close to the resistors. Any thermal switch whose contacts open at 80 °C can be used.

The current drawn from the battery is only about 10 μA, which rises to around 10 mA when the buzzer is energized.

Design: H. Bonekamp [944001]

INDICATOR FOR LEAD-ACID BATTERY CHARGER

This unit is intended as an add-on for the many lead-acid battery chargers that have no charge indicator. A green LED shows that the battery is connected with correct polarity. A red LED indicates that the battery voltage has reached its operating voltage, that is, the battery is fully charged. A yellow LED functions as on/off indicator, that is, shows that the mains is connected.

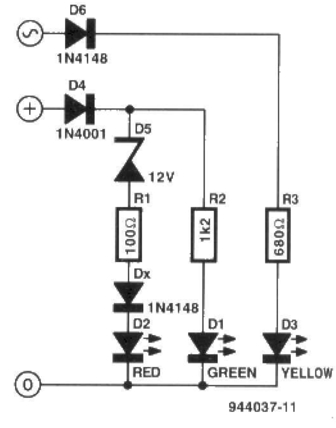
The operation of D_1 and D_3 is straightforward and needs no explanation. In the prototype, D_2 glowed faintly at 13.5 V and brightly at 14.4 V. Because of the tolerances in the zener and LED voltages, it may be necessary for a more positive indication to add another diode in series with D_x or even to omit D_x. Owing to the vast differences between commercial chargers, this can be ascertained only by trial and error.

Most standard chargers for 12 V lead-acid batteries provide a voltage of 13.8 V, while fast chargers provide 14.4 V.

Standard LEDs can be used.

The + and 0 terminals in the diagram must be connected to the corresponding terminals of the battery. The ~ terminal must be linked to the secondary transformer winding with good-quality cable. Make sure that it is connected to the secondary and not to the primary, because that could be fatal.

Design: K. Walraven [9440437]



80C451 CONTROLLER BOARD

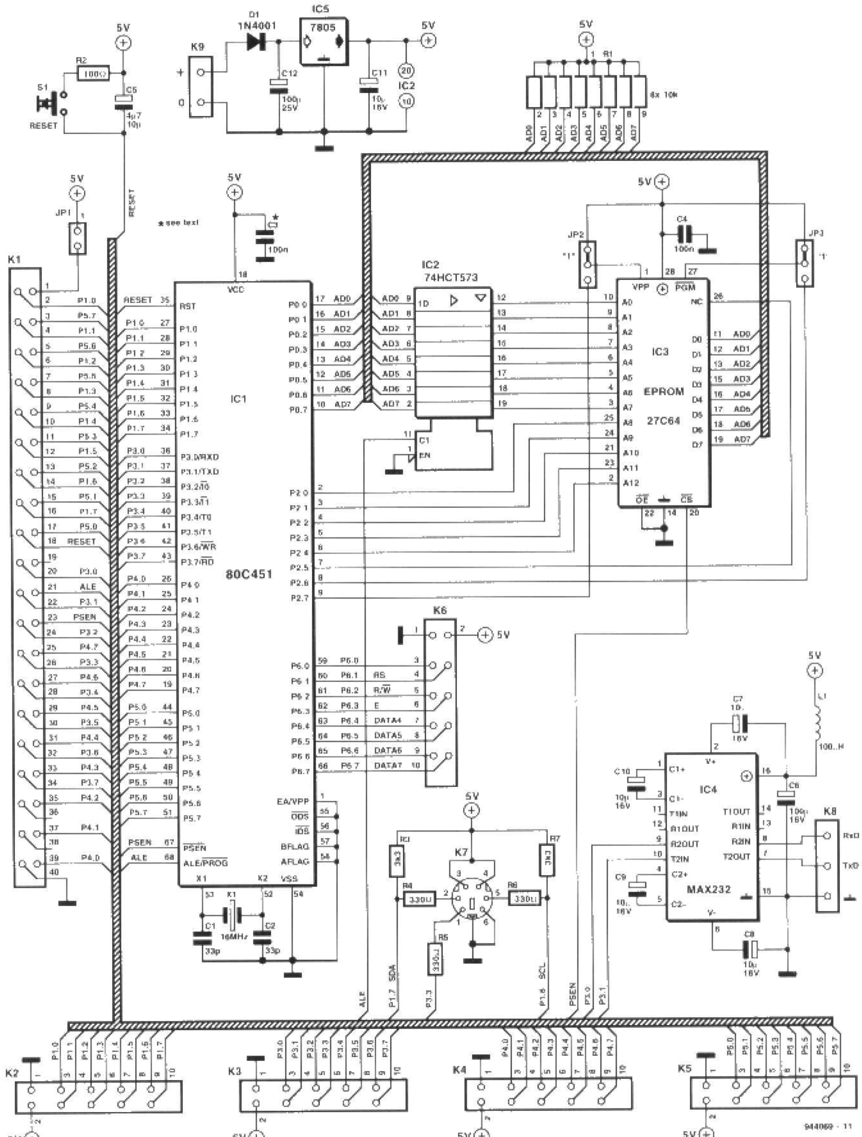
The 80C451 from Signetics is one of the many derivates of the 'generic' 8051 originally manufactured by Intel. The obsolescent SC80451CCN64 was first discussed in Ref. 1. The '451 has several ports more than the 8051. Here, a small controller board is presented based on the 68-pin PLCC version of the 80C451, the SC80C451CCA68 (12 MHz) or SC80C451CGA68 (16 MHz).

As indicated by the circuit diagram, the microcontroller board has all the 'typical' ingredients: memory formed by an EPROM, IC_3, an address latch, IC_2, a power supply, IC_5, a reset circuit, S_1-R_2-C_5, an RS232 interface, IC_4, and connectors (boxheaders) K_1-K_6, which allow you to hook up extension circuits. Perhaps less common is the on-board I^2C interface around mini-DIN connector K_7.

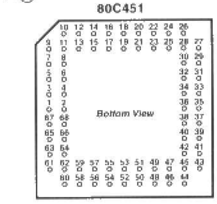
Boxheaders K_2-K_6 give access to the controller's plethora of port lines. Boxheader K_6 is wired to allow easy connection of an LCD display. Such a display can be used in 4-bit mode only, while the contrast adjustment pot must be located on the unit itself. Also note that the associated 'contrast' pin, number 3, of the LCD module is **not** connected to the controller board.

If a flatcable, a 40-pin DIP header and a 40-way IDC socket are used, boxheader K_1 gives a 1-to-1 correspondence with the pins of the 'standard' 8051 controller in a 40-pin DIL enclosure. However, to prevent problems, the '451 clock signal is not copied to the socket. Jumper JP_1 allows you to connect the 5-V supply of the '451 system to the 8051 system. The 40-way DIL socket enables the controller board to be turned into a simple 8751 emulator (see also Ref. 1). In that setup, port 0 of the emulated 8751 equals port 5 of the '451. Similarly, port 2 of the emulated 8751 then equals port 4 of the '451.

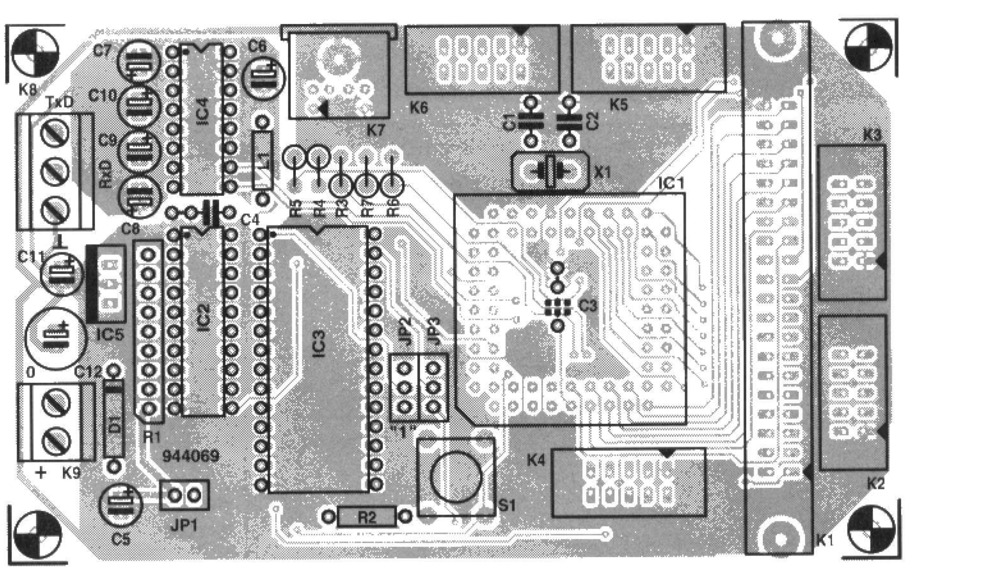
Although an I^2C socket is provided on the board, that should not be taken to mean that the 80C451 has built-in hardware to interface with an I^2C bus. It should be noted that the I^2C lines of the controller board



80C451



pin	27C64	27C128	27C256	27C512
1 (JP2)	Vcc	Vcc	Vcc	A15
26	xx	A13	A13	A13
27 (JP3)	Vcc	Vcc	A14	A14

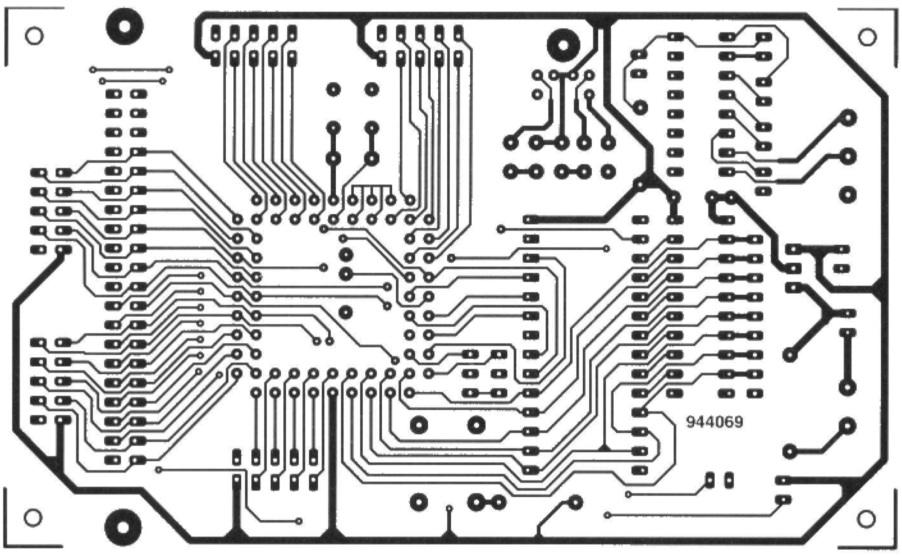


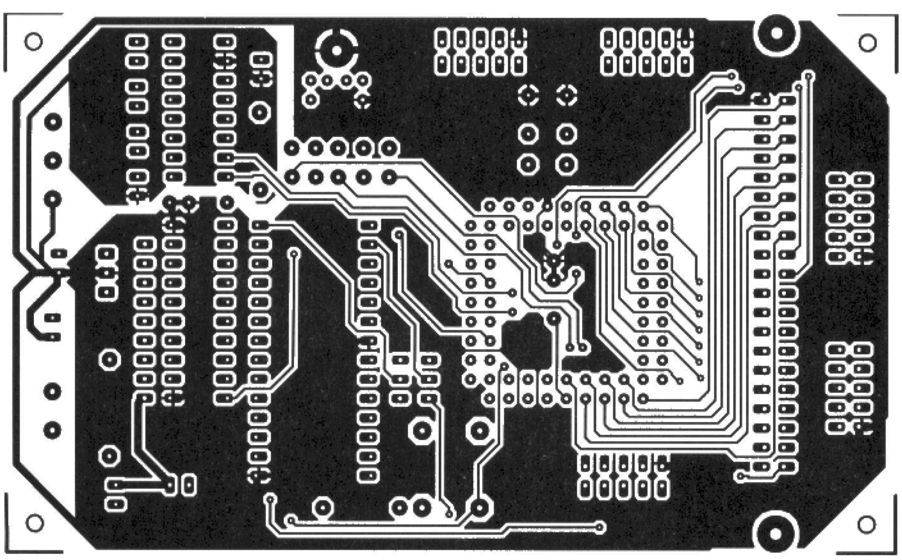
are capable of supplying up to 1.5 mA, which is less than a standard I^{2}C line (3 mA). However, this need not cause problems because 3.3-kΩ pull-up resistors are used here. On the same tack, the switching thresholds of the '451 inputs are not to I^{2}C standard, but no real problem there, either. Fortunately, a number of elementary software routines to implement I^{2}C communication using the 80C451 are available from the Philips Semiconductors databank (bulletin board) in Holland which can be contacted by modem on (+31) 40 721102. Dial up and download the file I2CBITS.EXE.

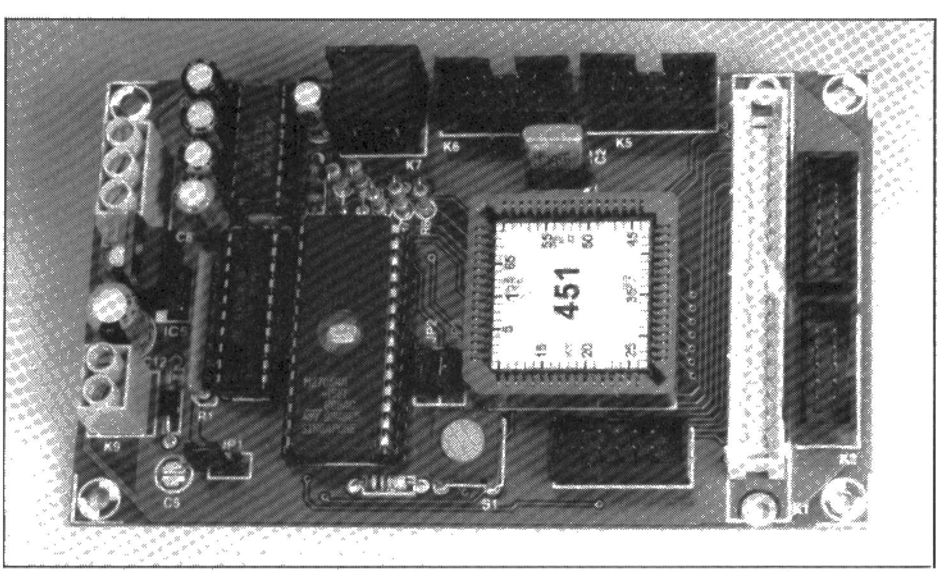
The RS232 channel is a standard application of the MAX232 single-chip RS232 interface. Although only RxD and TxD are implemented, this should work in most, if not all, cases where a simple link is desired to a PC running a terminal emulation program.

All of the popular EPROM types 2764 through 27512 may be used in this circuit — the selection is made with two jumpers, JP_2 and JP_3 (see table). Note that pin 26 (n.c. on the '64) is permanently tied to address line A13.

Finally, the quartz crystal frequency is 12 MHz or 16 MHz, depending on the controller type used. Capacitor C_3 is mounted at the underside of the board for maximum decoupling efficiency. Current consumption of the board is of the order of 30 mA, depending on clock speed and connected extensions. In practice, a 100 mA power supply will be adequate for all and sundry applications.







Parts list

Resistors:
R_1 = 8-way (9-pin) 10 kΩ SIL
R_2 = 100 Ω
R_3, R_7 = 3.3 kΩ
R_4–R_6 = 330 Ω

Capacitors:
C_1, C_2 = 33 pF
C_4 = 100 nF
C_3 = 100 nF (fit at underside of board)
C_5 = 4.7 µF, 10 V, radial
C_6 = 100 µF, 10 V, radial
C_7–C_{11} = 10 µF, 16 V, radial
C_{12} = 100 µF, 25 V, radial

Inductors:
L_1 = 100 µH choke

Semiconductors:
D_1 = 1N4001

Integrated circuits:
IC_1 = SC80C451CCA68 (12MHz) or SC80C451CGA68 (16MHz) (Signetics/Philips)
IC_2 = 74HCT573
IC_3 = 27C64 (see text)
IC_4 = MAX232
IC_5 = 7805

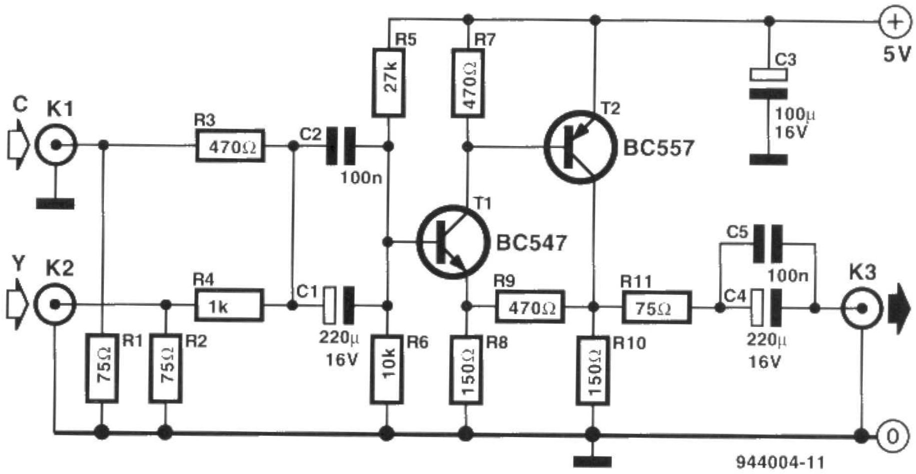
Miscellaneous:
JP_1 = 2-way jumper.
JP_2, JP_3 = 3-way jumper.
K_1 = 40-way boxheader.
K_2–K_6 = 10-way boxheader.
K_7 = 6-way mini DIN socket, PCB mount.
K_8 = 3-way PCB terminal block.
K_9 = 2-way PCB terminal block.
S_1 = push-button 2CTL2
X_1 = 16MHz or 12MHz crystal.
PCB Ref.944069-1 (p.110).

Reference:
1. 8751 emulator, *Elektor Electronics* March 1992.

Design: K.M. Walraven
[944069]

S-VHS-TO-VHS CONVERTER

Most modern video units are provided with an S-VHS or Y/C output. This output furnishes the black-and-white information (luminance or Y) and the colour information (chrominance or C) of a video picture on separate pins. This separation of data improves the quality of a picture generated by an S-VHS signal appreciably compared with a standard CVBS signal. Unfortunately, older television receivers without an S-VHS input can not process such a signal. It is for owners of such older receivers that the present circuit was designed: it recombines the Y and C components into a CVBS signal.



The two components of the S-VHS signal are applied to the converter via K_1 and K_2. The level of the luminance component is 1 V_{pp} and that of the chrominance component is 0.5 V_{pp}. For that reason, a weighting factor is applied in the recombination of the components. The output signal is composed of 1/3 of the luminance component and 2/3 of the chrominance component. The level of the signal at the base of T_1 is thus 666 mV. The amplification of the circuit is ×3 (R_9:R_8), so that the level of the signal at the collector of T_2 is about 2 V_{pp}. The potential divider consisting of R_{11} and the input impedance of the receiver (75 Ω) halve the signal being applied to the receiver, whose level is, therefore, 1 V_{pp} again.

If the input of the television receiver is adequately decoupled for d.c., capacitors C_4 and C_5, as well as R_{10}, may be omitted. This presupposes that the input impedance is 75 Ω for both the a.c. and the d.c. component.

The converter draws a current of about 25 mA.

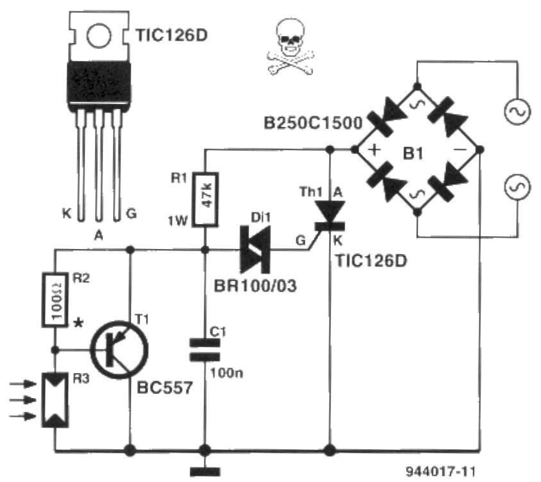
Design: J Kircher

[944004]

LIGHT-SENSITIVE SWITCH

The switch is typified by the very few components needed and its high sensitivity. Because of this, it has some limitations:

- It is suitable for use with incandescent lamps only.
- It has no switch-on delay, so that the lamp will light also during the day if the sensor becomes shaded.
- It has no switching threshold, so that the lamp will light up gradually.



The circuit is connected in series with the lamp. The bridge rectifier ensures that both period halves of the alternating voltage can be switched with the aid of thyristor Th_1. The gate of Th_1 is driven via R_1-C_1 and diac Di_1. At the onset of each period half, C_1 is charged rapidly via R_1. As soon as the potential across the capacitor has reached some tens of volts, the diac switches, whereupon Th_1 begins to conduct. It continues to do so until the next zero crossing, when it gets a fresh gate pulse, and so on.

When transistor T_1, which is in parallel with C_1, begins to conduct, the potential across C_1 remains virtually nil, so that the thyristor can not be ignited.

When the light-dependent resistor (LDR) between the base and collector of T_1 is in light, it has a low resistance so that T_1 is on. The thyristor, and thus the lamp, is off. When the LDR is in darkness, T_1 is off and the thyristor, and thus the lamp, is on. The value of resistor R_2 determines the moment of switch on, so it can be chosen to individual requirements.

The circuit is so sensitive that great care must be taken to ensure that the LDR can not receive any light from the lamp (if it did, a sort of oscillator effect would ensue).

The LDR should be a type that has a resistance of about 60 Ω when it is in bright light and of about 2 MΩ in darkness.

It is best to use a Type TIC126D thyristor, not a Type TIC106D, because this is so sensitive that the lamp would remain on. If a TIC106D must be used, its sensitivity can be reduced by soldering a 220 Ω resistor between gate and cathode.

Since the full mains voltage is present at various points in the circuit, it is essential that the switch is built into a well-insulated enclosure.

Design: J. Voûte.

[944017]

ROBUST A.F. POWER AMPLIFIER

This is a no-frills audio power amplifier based on inexpensive transistors. It is short-circuit protected, and has a maximum power output of the order of 50 W into 4 Ω. As shown by the circuit diagram, the amplifier is a classic push-pull class-B design.

To minimize the offset current which flows through feedback resistor R_{10}, zener diodes D_1 and D_2 should be matched for equal zener voltages. Similarly, transistors T_7 and T_8, the complementary pair BD139-BD140, should be matched for equal base-emitter voltages, U_{be}. They are fitted on a common heatsink to ensure that they are always at the same temperature. If T_7 and T_8 are not matched, or if D_1 and D_2 have different zener voltages, it may not be possible to compensate the offset voltage at the amplifier output despite adjustment of preset P_1.

To negate the effect of the input stage off-set variation, the feedback line is decoupled by two parallel connected bipolar electrolytic capacitors, C_6 and C_7.

Like T_7 and T_8, the input transistors, T_1 and T_2, must be thermally coupled. This can be achieved in a simple way by clamping the two transistors face-to-face with the aid of a small band of aluminium, copper or brass.

Because of their floating bias, the input transistors are sensitive to supply voltage fluctuations. Each transistor is, therefore, provied with its own regulator consisting of a current sources (T_3-T_4) and a zener diode (D_1-D_2). Note that the tolerance on the FETs and the zener diodes may well cause a deviation of up to ±1 V from the nominal supply voltage required, which is ±18 V.

Capacitors C_8 and C_9 in the cascade stages formed by T_2-T_6 and T_1-T_5 serve to minimize the adverse effect of the base-collector capacitances of T_1 and T_2. The base-collector junctions of transistors T_7 and T_8 are shunted by capacitors (C_{10}-C_{11}) because the BD139 and BD140 altough electrically complementary types do not have the same switching speed.

When the amplifier is first switched on, the voltage across R_{17} and R_{18} will settle at a certain value, and then rise slowly by about 0.15 V. This is normal and mainly owing to the simple design and the inevitable thermal effects in T_1-T_2 and T_7-T_8. This variation, however, calls for a good zener device to keep the quiescent current stable. In other words, the zener voltage between the bases of T_{13} and T_{14} must be independent of the current variation through T_9 and T_{10}. The quiescent current is adjusted with preset P_2. The 'super' zener formed by T_9 and T_{10} has an a.c. resistance which is about five times lower than that of a conventional one-transistor stabilizer. For obvious reasons, both transistors are fitted on the same heatsink as the power output transistors.

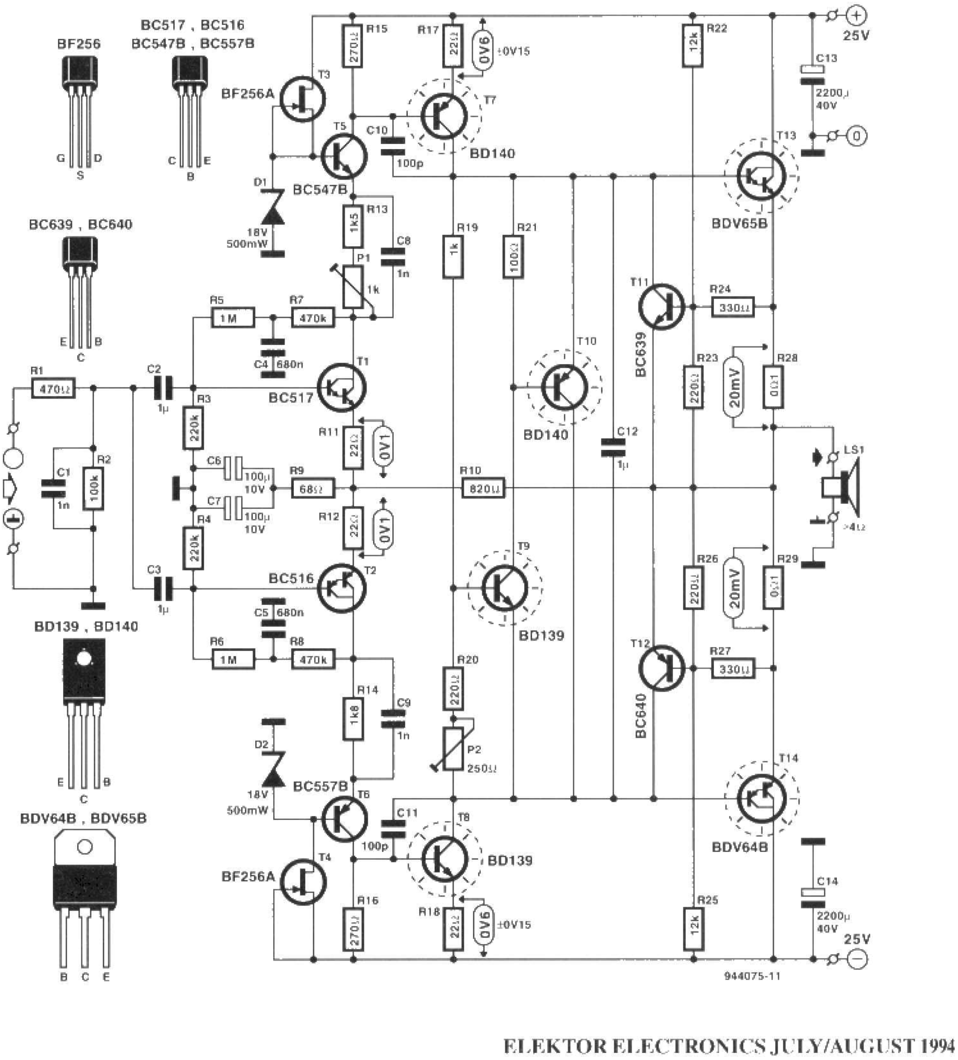
SOAR (safe operating area) protection is provided by transistors T_{11} and T_{12}. Resistors R_{22} through R_{27} have values which enable the currents through T_{13} and T_{14} to be kept within reasonable limits when the amplifier output is short-circuited or connected to a too low impedance.

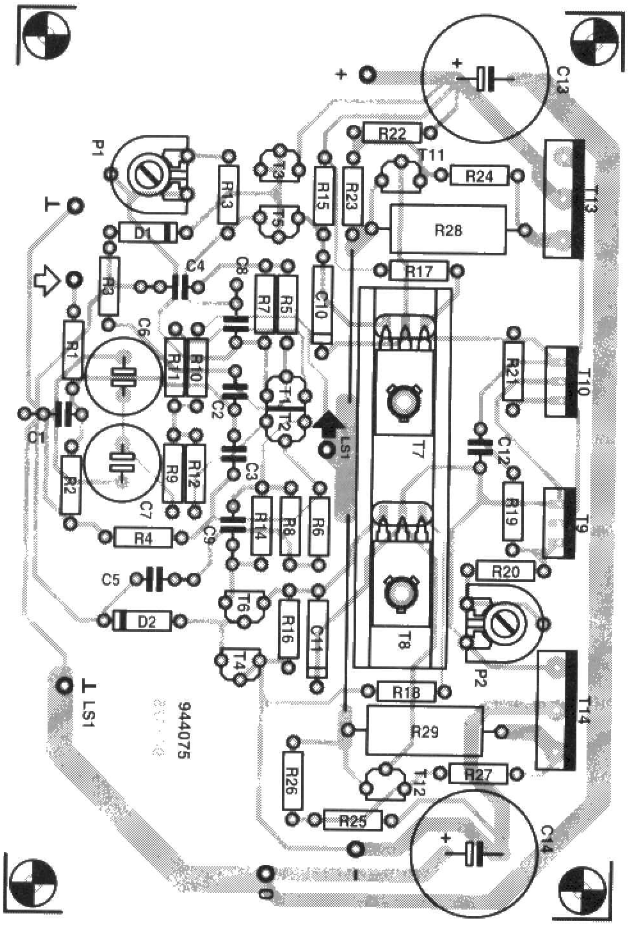
Main parameters

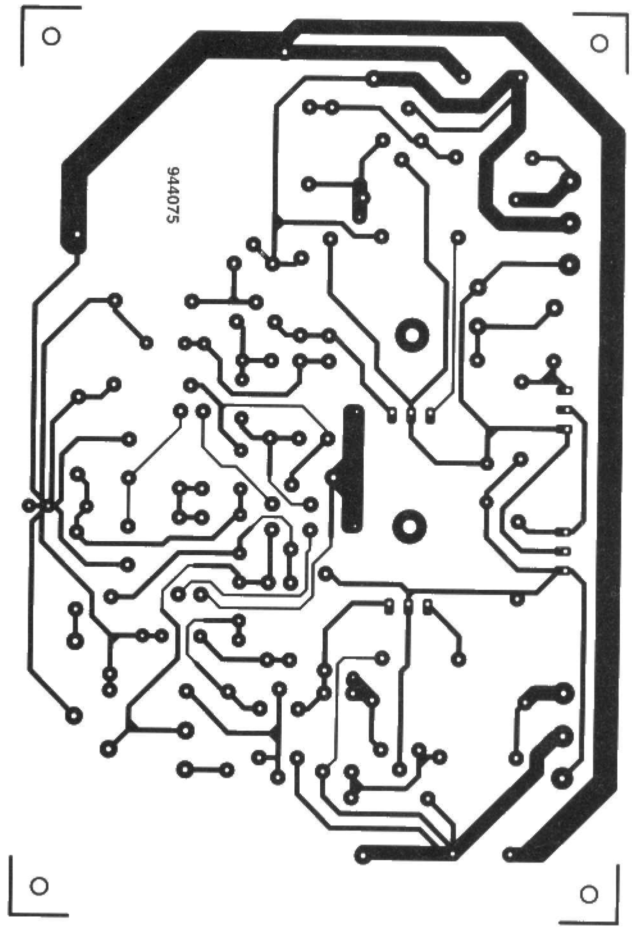
Input impedance:	47 kΩ
Input sensitivity:	1.25 V (30 W/8 Ω)
C1 not fitted:rise time	<0.7 μs
slew rate:	>40 Vμs
C1 fitted:	<1.5 μs
slew rate:	>24 μs
Bandwidth:	(30 W/8 Ω): 10 Hz - 180 kHz
THD+N: 1 W/8 Ω:	<0.005% (1 kHz)
25 W/8 Ω:	<0.02% (1 kHz)
25 W/8 Ω:	<0.07% (20 Hz - 20 kHz)
S/N (1 W/8 Ω):	>100 dB (B = 22 kHz)
P_{max} (THD+N = 0.1%):	30 W into 8 Ω
	56 W into 4 Ω
Damping (20 Hz - 20kHz):	8 Ω):>350

Measurements conditions:

Amplifier powered by a regulated ±25-V supply. Quiescent current (T13/T14) set to 200 mA; bandwidth 10 Hz to 80 kHz unless otherwise stated.







The construction of the amplifier should be evident from the circuit board layout and the photograph. Transistors T_9, T_{10}, T_{13} and T_{14} must be fitted with ceramic washers to keep their metal tabs isolated while still maintaining a low thermal resistance to the SK85 heatsink. T_7 and T_8 must also be fitted with washers (mica types are o.k.) on their common heatsink.

Be sure to set P_2 for maximum resistance (wiper fully counter-clockwise) before applying the supply voltage. Next, carefully adjust P_2 for a quiescent current of about 200 mA through T_{13} and T_{14}, which should correspond to about 200 mV across R_{28} and R_{29}. Finally, the minimum loudspeaker impedance that can be used with the amplifier is 4 Ω.

Parts list

(One channel)

Resistors:

R_1 = 470 Ω
R_2 = 100 kΩ
R_3, R_4 = 220 kΩ
R_5, R_6 = 1 MΩ
R_7,R_8 = 470 kΩ
R_9 = 68 Ω
R_{10} = 820 Ω
R_{11}, R_{12}, R_{17}, R_{18} = 22 Ω
R_{13} = 1.5 kΩ
R_{14} = 1.8 kΩ
R_{15}, R_{16} = 270 Ω
R_{19} = 1 kΩ
R_{20}, R_{23}, R_{26} = 220 Ω
R_{21} = 100 Ω
R_{22}, R_{25} = 12 kΩ
R_{24}, R_{27} = 330 Ω
R_{28}, R_{29} = 0.1 Ω, 5 W
P_1 = 1 kΩ preset H
P_2 = 250 Ω preset H

Capacitors:

C_1, C_8, C_9 = 1nF
C_2, C_3, C_{12} = 1 µF, pitch 5 mm
C_4, C_5 = 680 nF
C_6, C_7 = 100 µF, 10 V, bipolar, radial
C_{10},C_{11} = 100 pF, 160 V polystyrene
C_{13}, C_{14} = 2200 µF, 40 V, radial

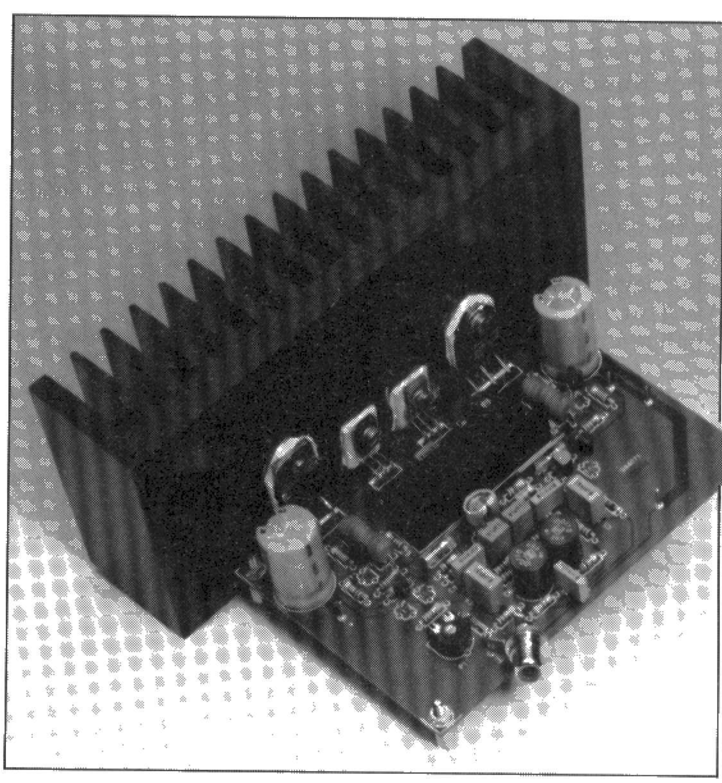
Semiconductors:

D_1,D_2 = 18 V zener, 0.5 W
T_1 = BC517
T_2 = BC516
T_3, T_4 = BF256A
T_5 = BC547B
T_6 = BC557B
T_7, T_{10} = BD140
T_8, T_9 = BD139
T_{11} = BC639
T_{12} = BC640
T_{13} = BDV65B
T_{14} = BDV64B

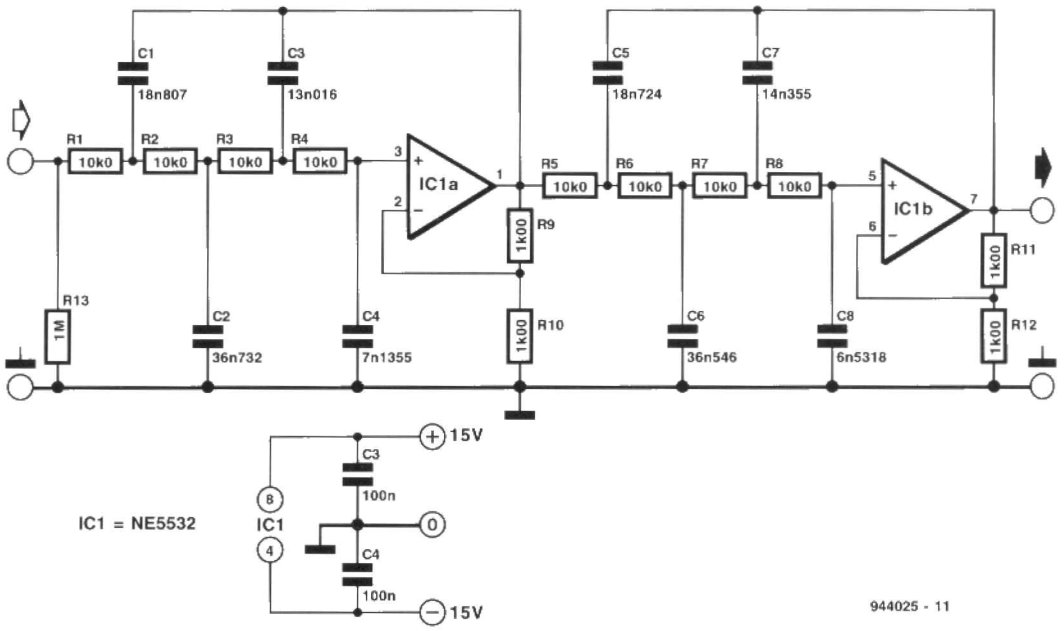
Miscellaneous:

Common heatsink for T7-T8 Type SK12/50 mm (15 K W^{-1}; Fischer).
2 TO-220 style ceramic washers.
2 TOP-3 style ceramic washers.
2 TO-220 style mica washers (cut to size).
Heatsink 0.65 K W^{-1} for T9–T10, T13–T14, e.g, SK85/75 mm (Fischer).
PCB Ref. 944075 (p. 110) (two required for a stereo amplifier).

Design: T. Giesberts
[944075]



8TH ORDER BUTTERWORTH FILTER



As usual in *RC* filter designs, resistors R_1–R_4 and R_5–R_8 are given equal values. Unfortunately, again as usual, this leads to rather awkward values of the capacitors. However, rounding off these values to the nearest E12 value leads to ripples in the transfer characteristic. The specified component values give a cut-off frequency of 1 kHz.

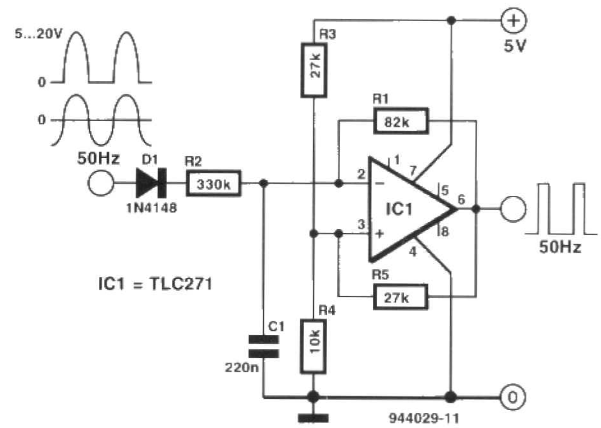
The frequency characteristic is rather different from that of a Bessel filter, as is the transit time. Moreover, there is a tendency to ringing. This is because the second section has a 3 dB peak in its gain just before the cut-off point. However, in practice this is hardly noticeable: the prototype was found to be usable up to full drive of the opamps. The only noticeable effect was a (very) slight deterioration in the signal-to-noise figure around the cut-off frequency.

Although the NE5532 proved well up to the task, it may be worthwhile to use opamps with FET inputs. These generate rather more noise than bipolar types, but, since the noise in the filter is caused mainly by the resistor, this would not matter much.

Each of the NE5532 chips draws a current of about 4 mA.

Design: T. Giesberts [944025]

MAINS-SYNCHRONISED OSCILLATOR



The mains-synchronised oscillator has certain advantages over a simple zero crossing detector. For instance, the brief failing of input pulses does not immediately cause a disaster and spurious pulses on the mains have hardly any effect on the circuit.

Because of the feedback to the positive input of IC_1 via R_5, the opamp has some hysteresis. This causes the output to change state when the potential across C_1 exceeds the upper hysteresis threshold and to change state again when that voltage drops below the lower hysteresis threshold. Since C_1 is continuously charged via R_1, the output of IC_1 continuously changes from one state to the other, so that the opamp behaves like a rectangular-signal generator. The duty factor depends on the threshold voltage; with values of R_3 and R_4 as specified, it is not greater than 50%. The frequency of the oscillations is determined mainly by R_1 and C_1 and should be just a little higher than the mains frequency (55–60 Hz).

The oscillator is synchronized with the mains frequency by connecting the anode of D_1 to the secondary winding of the supply transformer (that is, prior to the bridge rectifier). The positive pulsating direct voltage causes the discharge time of C_1 to lenghten, so that the oscillator frequency drops and synchronises with the mains frequency. The synchronisation causes a phase shift that depends on the strength of the input signal and on the difference between the mains and oscillator frequencies.

If the level of the 50 Hz input signal is <5 V or >20 V, the value of R_2 should be changed accordingly.

Before the circuit is taken into use, check the frequency of the freewheeling oscillator (input shorted to earth): this must be slightly higher than 50 Hz. The frequency depends to some extent on the opamp used.

Suitable opamps are the types LM358, TLC272, TLC271 and the TLC2201; note that a 741 is **not** suitable. If fast edges at the output are desirable, an LM339 is a good choice. Note, however, that this type has an open-collector output, so that a resistor of a few kΩ must be added between the output and the positive supply line.

The circuit as shown draws a current of only a few mA.

Design: K. Walraven [944029]

JOYSTICK-TO-MOUSE ADAPTOR

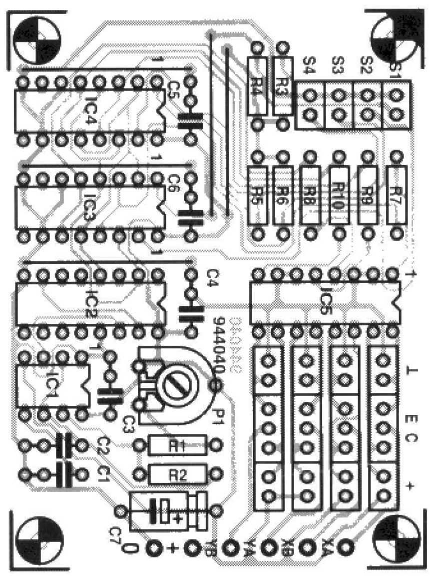
Although most PC games and more serious programs are perfectly suited to mouse control, there are applications, such as flight simulators, where joystick control would give a far more realistic 'feel'. Unfortunately, not all programs where joystick control would be desirable actually support such a device. In these cases, only mouse control is available. Also, in more general terms, the following problems are often encountered when running flight simulators:

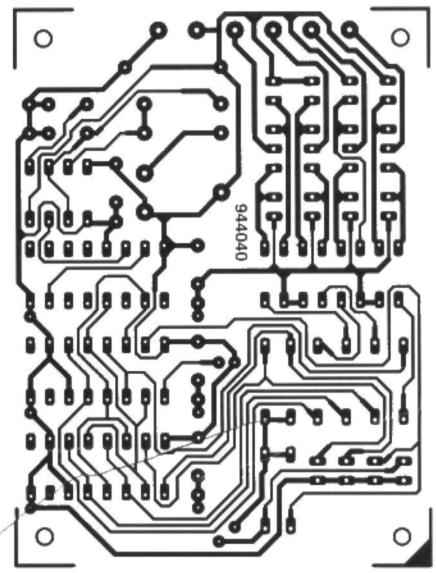
- although the resolution of the mouse is adequate for accurate control, the experience of flying is not simulated;
- the analogue joystick is far too inaccurate, and is pro-cessed digitally in any case;
- the digital joystick with D-A converter in principle simulates full deflection of the analogue joystick.

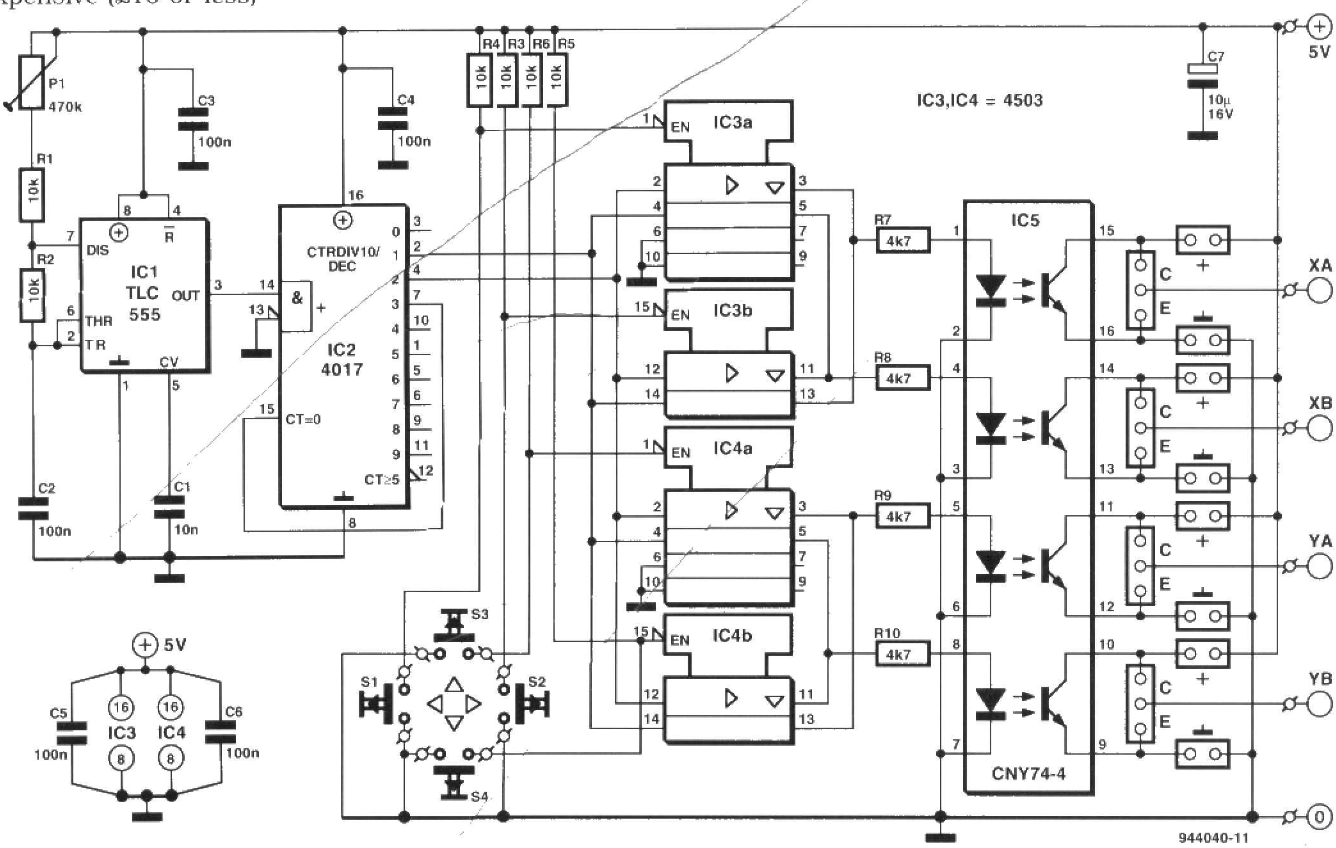
The circuit shown here allows a digital (ex-Commodore 64) joystick to be used for playing games that support mouse control only. The interface is based on electronics salvaged from an inexpensive (£10 or less) opto-mechanical mouse. The main board, the LEDs (or their series resistors) and the phototransistors are removed from the mouse. The phototransistors are replaced (electrically, that is) by four optoisolators contained in IC_5. This is achieved by connecting points XA, AB, YA and YB to the emitter or collector of each phototransistor removed. The adaptor board shown here has a number of jumpers to allow the collector and emitter of each phototransistor in the CNY74-4 to be wired to +5 V or ground. Jumpers E/C therefore determine whether the drive signal comes from the emitter or from the collector. The exact connections will depend on the electronics available on the mouse.

The circuit uses a TLC555 astable multivibrator and a 4017 divider to generate a pulse train which simulates the rotating slotted disc in the opto-mechanical mouse. This signal is routed to the optoisolators via two 4503 buffers. The desired direction (up/down; left/right) is selected with the digital joystick, via the enable inputs on the buffers. The speed of the cursor on the screen is set with preset P_1, since that determines the clock frequency of the TLC555.

The supply voltage for the interface is taken from the mouse, and can be found where the LEDs and their series resistors used to be connected. The







mouse/fire keys are wired directly, while the second 'fire' key (which is not available on the digital joystick) may be integrated into the adaptor enclosure.

Parts list

Resistors:
R_1–R_6 = 10 kΩ
R_7–R_{10} = 4.7 kΩ
P_1 = 470 kΩ

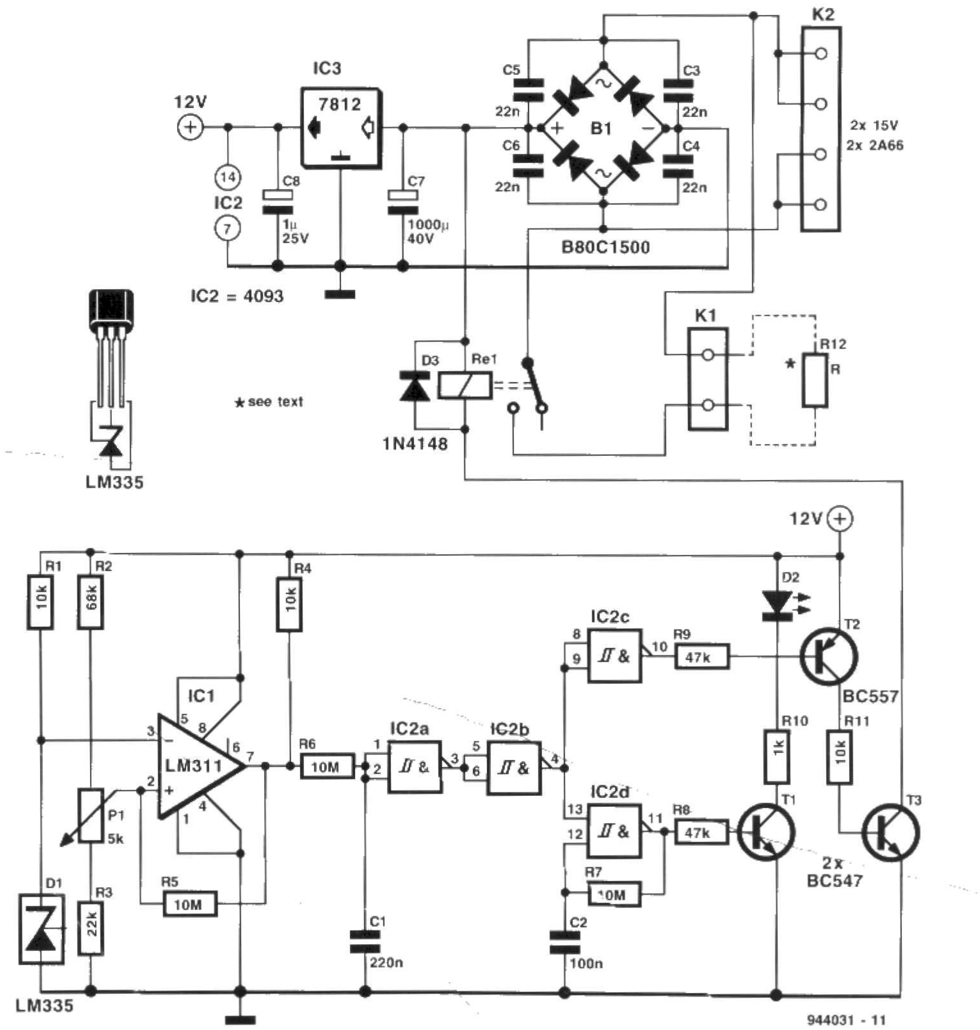
Capacitors:
C_1 = 10 nF
C_2–C_6 = 100 nF
C_7 = 10 µF, 16V

Integrated circuits:
IC_1 = TLC555
IC_2 = 4017
IC_3, IC4 = 4503
IC_5 = CNY74-4

Miscellaneous:
S_1–S_4 = switch in C64 joystick.
PCB Ref.944040 (p. 110).

Design: C. Wolff
[944040]

THERMO-CONTROLLED HOT-PLATE



The hot-plate was designed to keep a tray with developer for print material warm. It may also prove useful in a photographer's dark-room. It is contained in a small metal case, which also contains the electronics. The plate, which is made of aluminium and measures about 25×15cm (10×6 in) is mounted about 2 cm above the mains transformer. Four parallel-connected 22 Ω, 25 W resistors screwed to the underside of the plate function as heating element. The current through the resistors is switched on an off by an electronically controlled relay, which is also fitted in the metal case.

A Type LM335 temperature sensor is connected to the negative input of comparator IC_1. The voltage across the sensor is directly proportional with the temperature, so that when the hot-plate cools off, it will at a given moment drop below the reference voltage set by P_1 at the positive input of IC_1. The output of the comparator will then go high, so that after three inversions in IC_{2a}, IC_{2b} and IC_{2c}, there is a low logic level at the base of T_2. Since this is a p-n-p transistor, it will begin to conduct, whereupon T_3 is also switched on and relay Re_1 is energized. The heating resistors (represented by R_{12}), connected to K_1, are then connected to the mains transformer via K_2 and heat the hot-plate.

When the temperature causes the voltage across D_1 to exceed the reference voltage at pin 2 of IC_1, the comparator output goes low and T_2 and T_3 are switched off. The hysteresis provided by R_5 ensures that the on and off thresholds are far enough apart to prevent relay clatter.

An optional indicator is formed by IC_{2d}, T_1 and D_2. When the heating resistors are not switched on, the comparator output is low. Pin 13 of IC_{2d} is then also low and the gate functions as an inverter, so that T_1 is switched on and D_2 lights. When the heating resistors are on, pin 13 of IC_{2d} is high. Network R_7-C_2 then causes the gate to function as a rectangular-signal generator so that D_2 flashes in sync with it.

In the prototype, a toroidal mains transformer from ILP (2×15 V, 2×2.66 A) was used.

The power resistors were aluminium types provided with fixing holes.

The relay was a 24 V type from Siemens that can switch currents of 6 A or greater.

Design: R. Lucassen
[944031]

PC OVER-TEMPERATURE ALARM

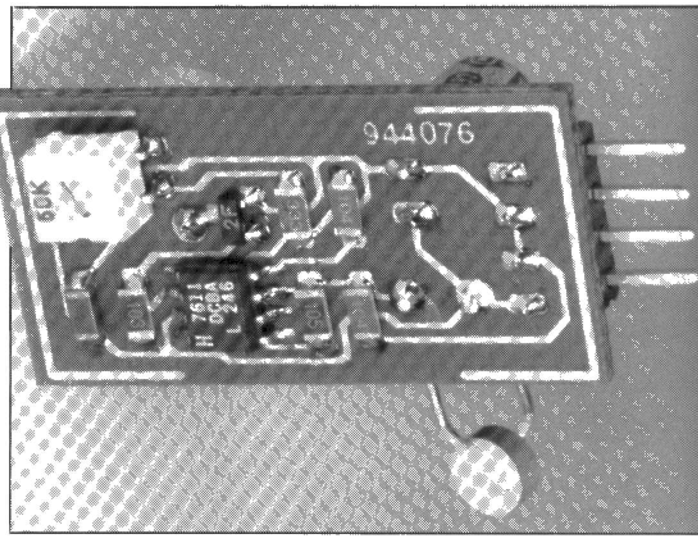
Any PC, however old, is too valuable to break down as a result of inadequate cooling, usually caused by fan failure. The alarm causes a buzzer to sound when the temperature inside the PC reaches a predefined level. Obviously, this early warning should prompt you to switch off and take a very serious look at the cooling of the PC if you do not want it to be turned into a lot of silicon junk. You may have fitted too many insertion cards, or the fan has failed. In any case, the cost of the present circuit is always lower than that of a new motherboard or a power supply.

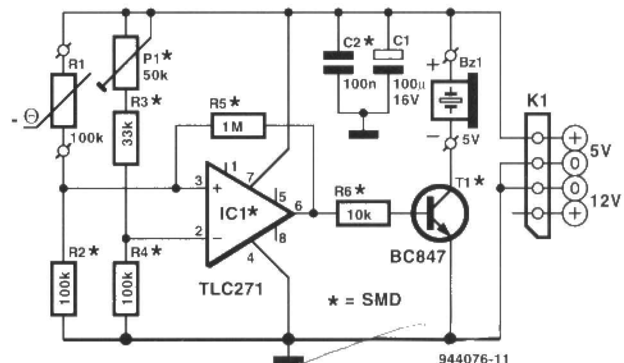
The temperature sensor used is an NTC (negative temperature coefficient) resistor, which is fitted at a suitable location in the air flow being maintained by the fan inside the power supply. The NTC is connected to a comparator whose output swings high if the resistance value of the NTC is smaller than the sum of preset P_1 and fixed resistor R_3. The switching temperatures are29 °C or 50 °C (84 °F or 120 °F) with the preset set to maximum or minimum resistance respectively.

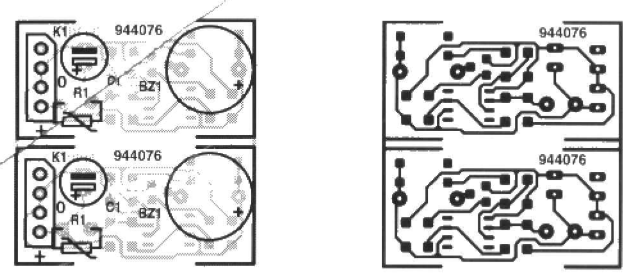
The alarm is adjusted by first turning P_1 to minimum resistance (wiper electrically towards R_3), heating the NTC to the desired alarm temperature (approx. 45 °C or 113 °F as a guide), and then adjusting P_1 until the buzzer just sounds.

The board is cut into two to enable two alarms to be built. The SMDs (surface mount devices) are fitted at the copper side of the board, and the conventional parts at the top side, as shown by the component overlays.

The circuit is powered via a 3.5-inch drive connector, which can be connected either way to the board without affecting the operation or the temperature range of the alarm. The NTC may be mounted directly on to the board, or off the board at a suitable location in the PC. In the latter case, the device is connected to the board via two short wires.







Parts list

Resistors:
R_1 = NTC 100 kΩ, Siemens series K164, B=4600K, R_2,
R_4 = 100 kΩ, SMT
R_3 = 33 kΩ, SMT
R_5 = 1 MΩ, SMT
R_6 = 10 kΩ, SMT
P_1 = 50 (47) kΩ preset (Bourns)

Capacitors:
C_1 = 100 µF, 16 V, radial
C_2 = 100 nF, SMT

Semiconductors:
T_1 = BC847B, SMT

Integrated circuits:
IC_1 = TLC271CD, SMT

Miscellaneous:
K_1 = 4-way SIL connector (3.5" drive supply).
Bz_1 = 5V DC buzzer
PCB Ref. 944076 (p. 110).

Design: L. Lemmens
[944076]

'NEAR-PERFECT' INPUT STAGE

As far as audio enthusiasts are concerned, striving for perfection is a never-ending quest. The present circuit should appeal to them.

The designer of amplifiers aims for the best possible symmetry, because that provides the optimum performance. Input stages often consist of two differential amplifiers that are each other's complement.

The d.c. setting of each differential amplifier is effected by a single-transistor current source, whose reference is a zener diode or LED. In itself, this is an excellent design, particularly when an LED is used, because this has a temperature coefficient virtually identical to that of the transistor. However, in practice, these values will hardly ever be exactly the same. Since, moreover, the thermal coupling normally leaves something to be desired, it is clear that there will be variations in the d.c. setting in practice. Furthermore, as each differential amplifier has its own current source, these adverse effects are doubled. The inevitable result is that thetwo stages do not operate in perfect symmetry.

In the present circuit, the differential amplifiers have a common current source. Any variations in this have a symmetrical effect, producing as it were automatic compensation. This means that the d.c. operating point is far more stable and the offset drift caused by temperature variations is much smaller. The only factor that still causes troubles is that

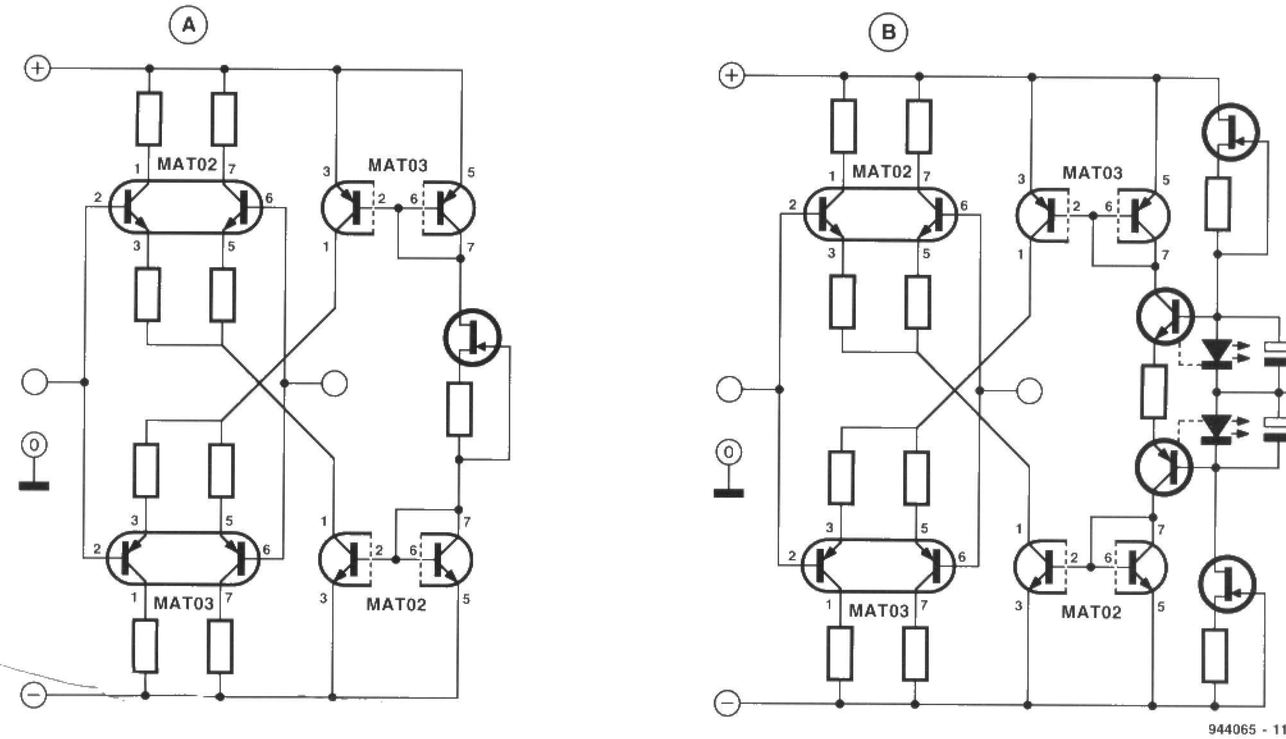
there are no true complementary dual transistors on the market. Even in the MAT02 and MAT03, the h_{FE}s are too far apart, but better types are not yet available.

The 'near-perfect' input stage in its simplest form is shown in diagram A. The current source, built from a JFET is enclosed by two current mirrors. The upper mirror fucntions as the current source for the lower differential amplifier and vice versa. The stability of the JFET source can be improved as shown in diagram B. Here the JFET is replaced by two transistors with a common emitter resistance. Two LEDs provide the setting of the base voltage; the current through these diodes is held constant by two JFET current sources. The current mirrors are built from monolithic dual transistors.

The collector-emitter voltage of the MAT03 must not exceed 36 V.

The current drain of circuit A is three times the current source setting. That of circuit B is the same plus the current drawn by the LEDs.

Design: T. Giesberts
[944065]



POWER SUPPLY FOR LCD MODULE

Building an LCD module into an equipment (for instance, a voltmeter into a power supply) is often not straightforward, because its supply can not be taken from the supply of the equipment (perhaps because floating voltages are to be measured).

This difficulty is resolved by the present circuit, which derives a direct voltage from the existing d.c. supply and keeps it isolated with the aid of capacitors. This enables the LCD module to measure floating direct voltages.

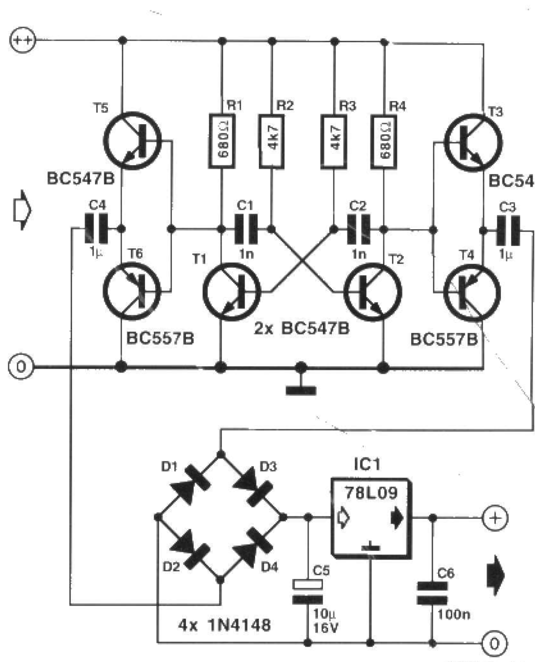
The circuit is based on an astable multivibrator, T_1-T_2, in a traditional configuration. The alternative charging and discharging of capacitors C_1 and C_2 will cause T_1 and T_2 to be switched on and off in turn. This results in rectangular, anti-phase pulses at the collectors of these transistors. These pulses are applied to T_3-T_4 and T_5-T_6 respectively.

The signals at the emitters of these buffer stages are applied to a bridge rectifier via C_3 and C_4 respectively. The direct voltage taken from the bridge rectifier is smoothed by C_5 and stabilized at 9 V by regulator IC_1.

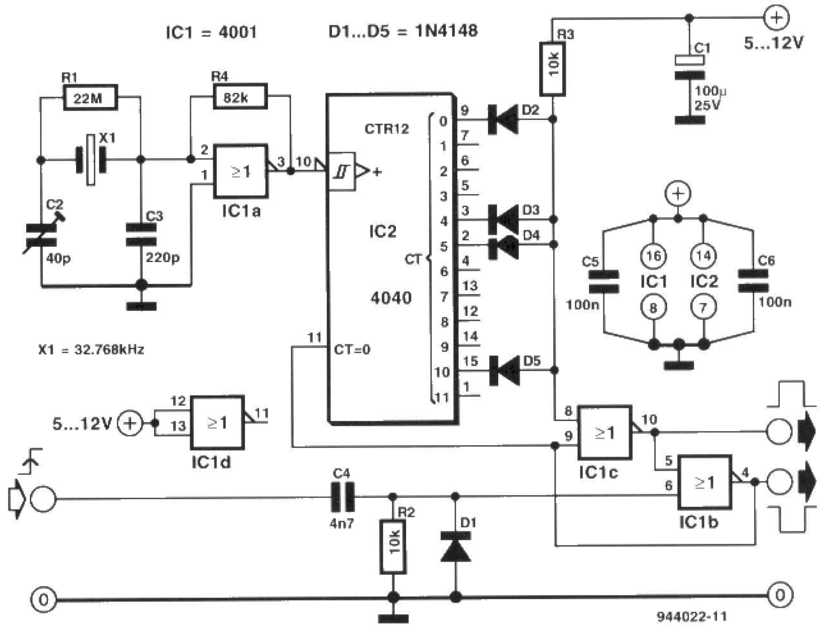
The current that can be drawn from the circuit is only a few mA, but this is sufficient for most LCD modules.

The input voltage must be in the range 12–15 V. If it is lower (8–12 V), IC_1 should be a Type 7806, assuming that the LCD module used can work from 6 V.

Design: R. Baltissen
[944039]



PRECISION ONE SHOT PULSE GENERATOR



This simple non-retriggerable pulse generator ('one-shot') is designed around an inexpensive digital watch crystal and two commonly available CMOS integrated circuits. Compared with the ubiquitous 74121/123 'external *RC*' mo-nostables and their derivates, this design is remarkably immune against supply voltage and ambient temperature variations. The output pulse width is defined with the aid of a 12-stage binary counter Type CD4040, at a precision equal to the period, T_c, of the crystal frequency.

NOR gate IC_{1a} is wired as a crystal oscillator, while IC_{1b} and IC_{1c} form a bistable. Diodes at the outputs of the counter provide a logic AND function. The bistable is set by a low-to-high transition at the trigger input of the circuit, whereupon the counter is enabled. When the counter reaches the count at which all cathodes of the diodes connected to the Q outputs of the counter are at logic high, the bistable and the counter itself are reset. Complementary one-shot pulses are available at the two outputs of the generator.

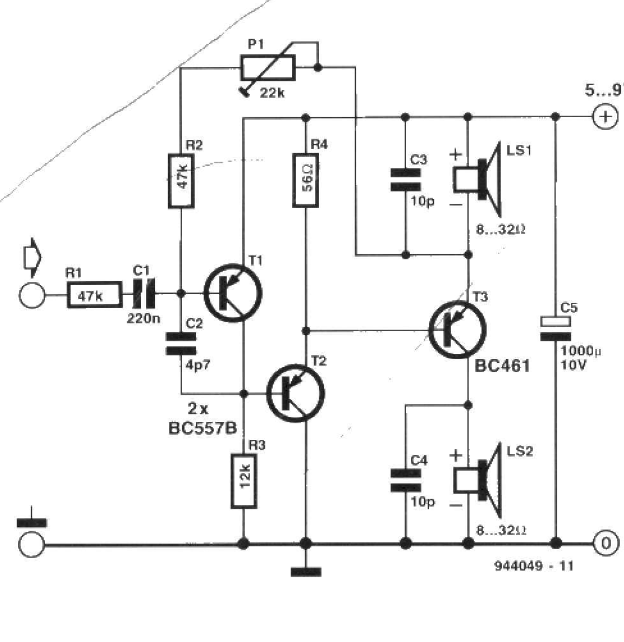
Up to 12 diodes may be connected to the counter outputs to program the divisor, *n*, that determines the width T_0, of the complementary output pulses:

$$n = T_0 \,/\, T_c.$$

A crystal frequency of 32.768 kHz will be one of the most cost effective, and results in a pulse width resolution of 30.5 µs and a maximum pulse width of 124.9 ms. Alternatively, crystals of higher frequencies up to 10 MHz may be used to increase the pulse width resolution. The generator works at a supply voltage range from 5 V to 15 V at a current consumption of a few milliamps, depending mostly on the quartz crystal frequency.

Design: M.S. Nagaraj
[944022]

TWO-SPEAKER AMPLIFIER



The design of this amplifier is based on the assumption that, since the collector current of a transistor is, roughly, the same as the emitter current, the transistor can drive two loudspeakers.

In the diagram, T_1 is a simple voltage amplifier, followed by emitter follower T_2 and class A output amplifier T_3. Negative feedback is provided by R_1-R_2-P_1. The amplification is about unity which can be altered slightly with P_1.

Noteworthy of the design is that the d.c. operating point of T_3 is determined by the base-emitter voltage of T_1. This voltage and the potential across LS_1 are virtually identical and the d.c. operating point of T_3 is thus practically independent of the supply voltage.

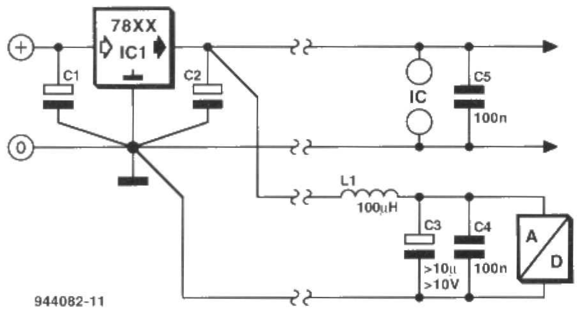
Inserting a second loudspeaker into the collector circuit of T_3 doubles the power output of the amplifier. The output power is 2×23 mW with a supply voltage of 5 V and 2×40 mW with a supply voltage of 9 V. In the latter case, distortion amounts to 0.1%. The frequency range extends from about 15 Hz to 200 kHz.

Since the feedback loop reference is the positive supply line, the power supply needs to be decoupled well (whence the high value of C_5).

The current drain depends primarily on the setting of T_3 and amounts to 100 mA at 5 V and 120 mA at 9 V.

Design: Amrit Bir Tiwana
[944049]

POWER SUPPLY DECOUPLING



In many advanced electronic circuits, speed and accuracy are of the utmost importance. It is sometimes overlooked that to achieve these, good decoupling of the power supply lines is an absolute must. 'Fast' ICs are particularly sensitive in this respect.

It can not be overemphasized that the connections of the decoupling elements to the supply pins of the IC must be as short as possible: every millimetre counts! Usually, the supply pins are located at opposite corners of the housing. As a rule of thumb, it may be taken that if the distance between these pins is doubled (owing to the layout of the PCB), the level of the resulting voltage variation is also doubled. That is significant.

Ideal, but expensive, are IC sockets with integral decoupling capacitor. It is, however, normally possible to achieve the same effect by soldering a small decoupling capacitor directly across the supply pins. It may not look nice, but it is effective. A value of 100 nF per (TTL, HC or HCT) IC is sufficient. Nowadays, there are capacitors designed specifically for decoupling on the market: the Sibatit series from Siemens is an example.

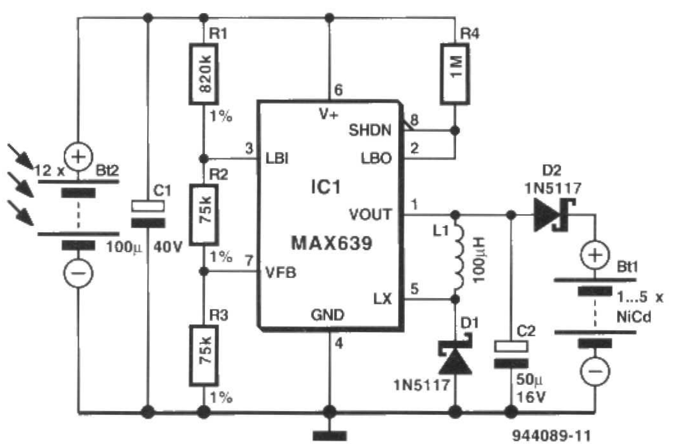
In circuits where the frequency is higher than 50 MHz, the effect of the decoupling is magnified when a 10 nF capacitor is soldered in parallel with the 100 nF capacitor.

A practical difficulty often arises in a hybrid circuit, that is, one with a digital section and an analogue section (e.g., an analogue-to-digital converter, ADC). Ideally, these sections should have separated supplies, but this is often not possible. If, therefore, only a single power supply is used, branch off a positive and a negative line for the analogue section immediately after the voltage regulator. In other words, **never** use combined supply lines in a hybrid circuit. As shown in the diagram, the effectiveness of the decoupling can be improved by inserting a resistance or inductance in the positive supply line. An inductance screens the analogue section from the spurious signals arising in the digital section.

It is essential that the entire analogue section is decoupled properly. This may, however, cause the inductance and capacitance to form a resonant circuit. This may be prevented by using a much larger capacitor and damping the circuit with a small series resistance. In practice, an *LC* filter consisting of a 100 µH inductor and a 10 µF capacitor works very satisfactorily. As shown in the diagram, the 10 µF capacitor is connected in parallel with the 100 nF capacitor (which remains essential!). A miniature choke (same size as a resistor) is perfect, because its internal resistance of 1–2 Ω is just right for damping the circuit.

Design: K.M. Walraven [944082]

SOLAR CELL-POWERED NiCd CHARGER



A battery of solar cells can charge a NiCd battery with better than 80% efficiency, provided the solar battery voltage exceeds the fully charged NiCd output by about 0.6 V. For that case, a simple blocking diode provides the charging path between the two batteries. Though inflexible, such dedicated systems are simple and effective.

The simple approach, however, is impractical if adjustment of the charging voltage is impossible or inconvenient. Badly mismatched voltages (<< 0.6 V) cause a low level of power transfer and consequent slow charging of the NiCd battery.

Adding a step-down switching regulator enables a given bank of solar cells to charge batteries of various terminal voltages at optimum rates and with efficiencies approaching that of the regulator itself.

The IC used for this purpose, the MAX639 from Maxim, operates in an unorthodox fashion, regulating the charging current such that the solar battery's voltage remains near the level required for peak power transfer.

The device therefore regulates its input voltage, rather than its output voltage as is customary. Potential divider R_2-R_3 disables the internal regulating loop by holding V_{FB} low. Voltage divider R_1-(R_2+R_3) enables the low battery input (LBI) to sense a decrease in the solar-battery voltage. Such decreases, which represent a move away from the solar cell's peak output power, cause the low battery output (LBO) to pull the SHDN input low and disable the chip. The LBI input then senses a rising input voltage, LBO goes high, and the resulting pulsating control maintains maximum power transfer to the NiCd battery. Current limiting in the IC limits the output current to 200 mA.

With the IC enabled, the regulator passes current from pin 6 to pin 5 through an internal switch resistance of less than 1 Ω. Combined with the regulator's low quiescent current (typically 10 µA) and high efficiency (typically 85%), this performance allows the circuit to deliver as much as four times the power of a single-diode circuit.

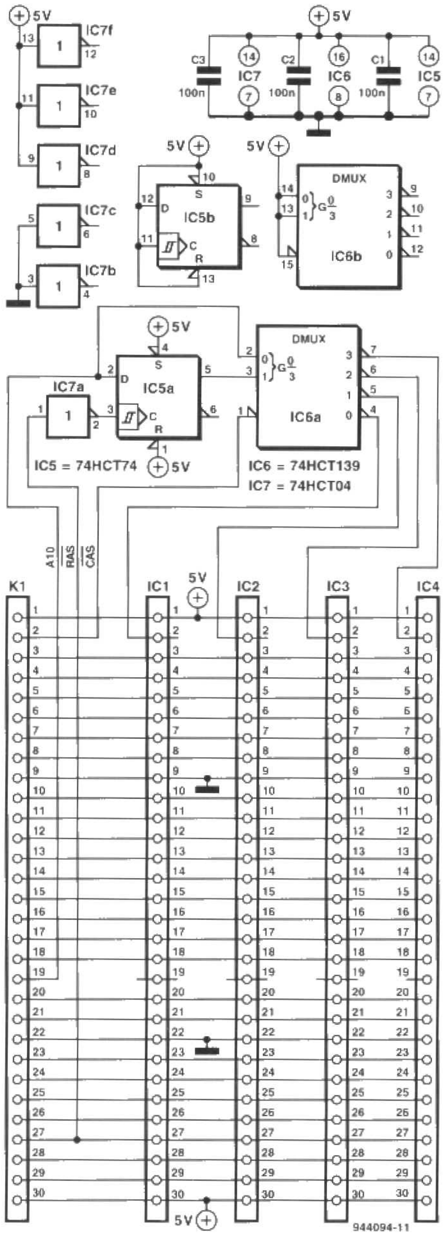
Note that the circuit should be used only to charge NiCd cells that can be charged continuously with a current of 200 mA, that is, have a capacity greater than 1700 mAh.

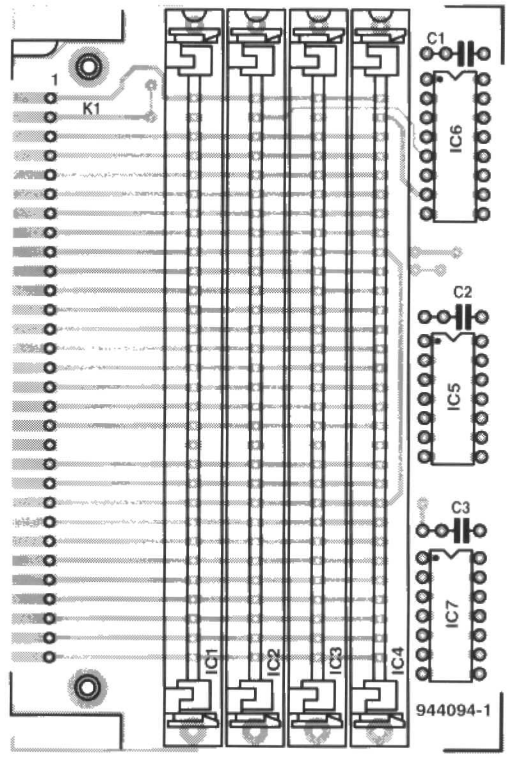
Maxim Application [944089]

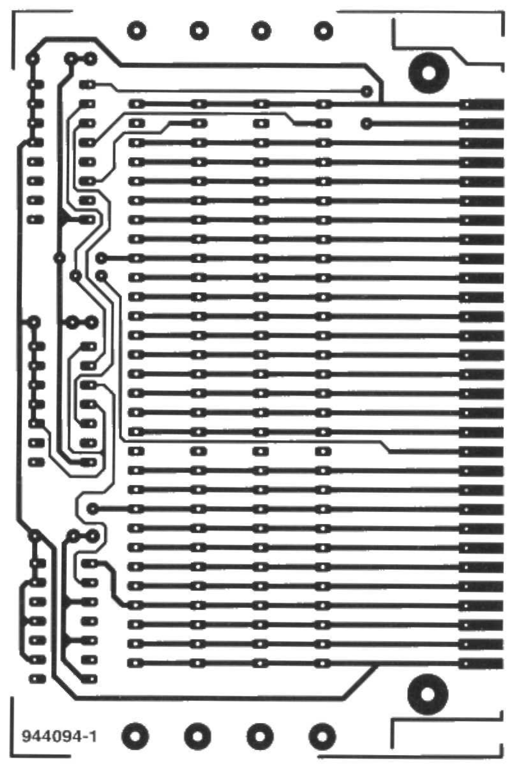
1–4 MBYTE SIMM ADAPTOR

Memory upgrades for PCs can present unexpected problems. For instance, if you are upgrading with 4-Mbyte SIMMs (single-in-line memory module), the existing 1-Mbyte SIMM will be useless in most cases. A waste of money? Not if you build the circuit shown here, which enables four 1-Mbyte SIMM modules with eight or nine DRAM chips, to act as a single 4-MByte SIMM, and occupy only one memory expansion socket. The circuit is **not** suitable for use with three-chip SIMMs.

The memory of a 486-based PC has a width of 32 bits, while a SIMM has a 'width' of only 8 bits (well, 9, if you include the parity bit). Consequently, memory modules can only be added four at a time. If 1-MByte SIMMs are used, that allows memories of 4, 8, 12 or 16 MByte to be created, or 16, 32, 48 or 64 MByte if you can afford to use 4-MByte modules. In some cases, 4-MByte and 1-MByte modules may be 'mixed', for instance, to give 20 MByte, consisting of four 4-MByte and four 1-MByte SIMMs. Thus, upgrading from 4-MByte to 16 MByte means that you have to exchange the four 1-MByte SIMMs with four 4-Mbyte modules. If you are lucky, the motherboard still has four sockets to insert your 1-MByte SIMMs, so that you can extend the memory up to 20 MByte. In other cases, for instance, when it is desired to upgrade a 12-MByte memory to 20 MByte, you may be left with eight 1-MByte modules for which there is no room on the motherboard.







Before deciding to build the present circuit, be sure to cast a look inside your PC to make sure that it has sufficient room in the memory expansion area for the circuit board and the SIMMs fitted on it. Particularly in mini-tower cases space may be tight, and the circuit can not be used.

Apart from the memory ICs used, the difference between a 1-MByte and a 4-MByte SIMM is that the latter has one more address line, A10. This line is used in conjunction with the two refresh signals RAS (row address select) and CAS (column address select) to select between the four 1-Mbyte SIMMs in the circuit. The level of A10 is latched in IC_{5a} (row address line A10). Together with column address line A10, it is decoded into a 1-of-4 selection signal for the respective SIMM module. This happens when the CAS signal is actuated. The contents of the SIMMs are refreshed by the briefly addressing all rows. Hence, all RAS connections on the SIMMs are interconnected.

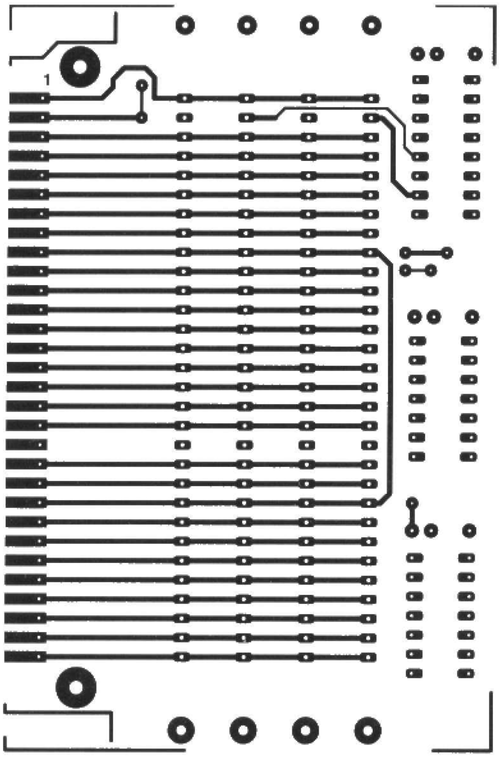
The adaptor board should be fitted with standard SIMM sockets, which can not be mounted the wrong way around because they are polarized. The SIMM module can also be fitted with the correct polarization only.

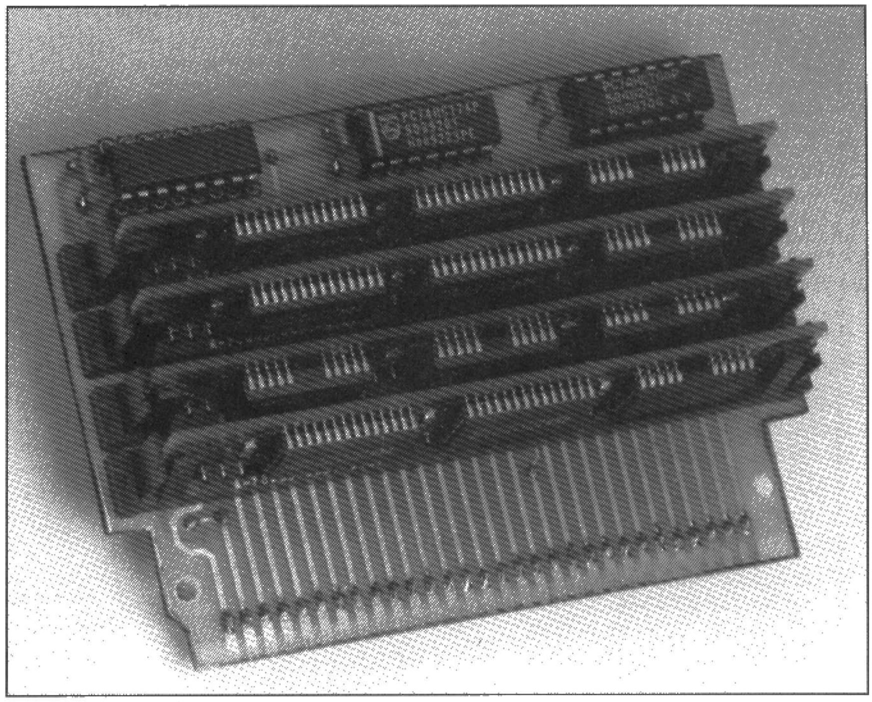
If you use more than one adaptor board, these can not normally be fitted alongside each other in adjacent slots. That is why the PCB is double-sided, allowing you to mount SIMM sockets at both sides, and fit adaptor boards next to each other. The ICs are, of course, fitted at one side only.

Parts list

Capacitors:
C_1, C_2, C_3 = 100 nF

Integrated circuits:
IC_1–IC_4 = SIMM socket, 30-way.
IC_5 = 74HCT74
IC_6 = 74HCT139
IC_7 = 74HCT04





Miscellaneous:
PCB Ref. 944094 (p. 110)

Design: A. Rietjens
[944094]

FREQUENCY INDICATOR

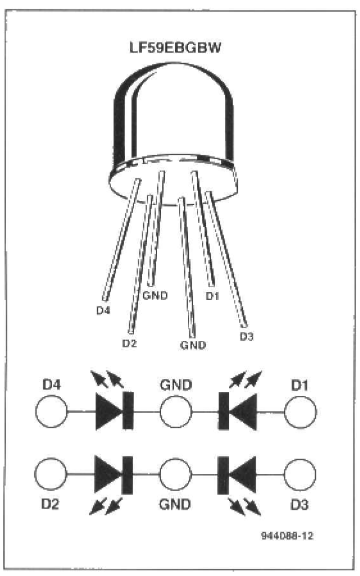
The circuit gives a coarse indication of the frequency by a tricolour LED, the Type LF-59EBGBW from Kingbright. When the frequency of the input signal is 50 Hz, the two blue LEDs light; when the frequency rises to 500 Hz, the red LED also lights, and when the frequency rises above 5 kHz, the green LED lights, too. The resulting hue is thus indicative of the frequency of the input signal.

The setting of P_1 determines the threshold of operation of the red LED. Since the green and blue diodes need a higher threshold voltage, the offset voltage of the red LED is increased by the static voltage amplification of the opamp.

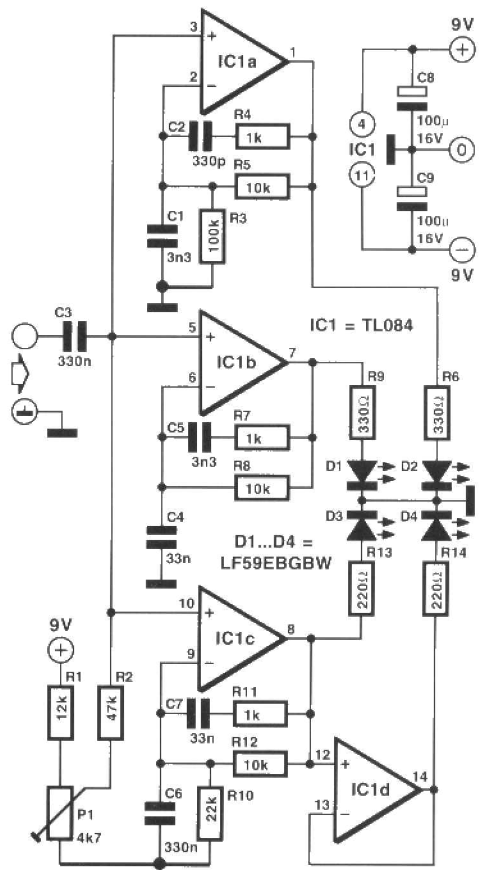
Networks in the feedback loops of the opamps provide cut-off points of 50 Hz (IC_{1c}); 500 Hz (IC_{1b}) and 5 kHz (IC_{1a}). These networks ensure frequency-independent brightness of the LEDs.

The stability of the opamps is enhanced by networks R_4-C_2, R_7-C_5 and R_{11}-C_7.

Since the brightness of blue LEDs is appreciable less than that of red or green types, two of them are fitted in the LF-59EBGBW. The second blue diode has its own buffer opamp, IC_{1d}. In spite of the fact that there are two blue diodes, these also draw twice as high a current as the red and green to give the same brightness. It is for that reason that the value of resistors R_{13} and R_{14} is lower than that of R_6 and R_9.



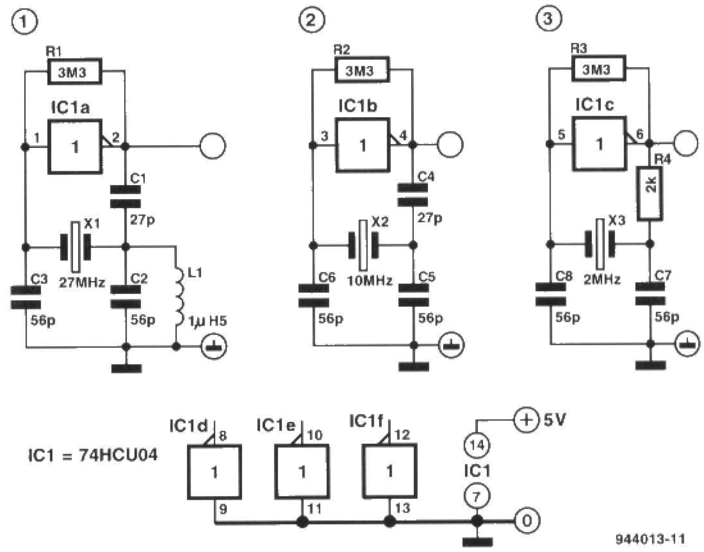
The level of the input signal should be 1 V_{pp}.



When all LEDs light, the circuit draws a current of about 100 mA.

Design: H. Bonekamp
[944088]

CRYSTAL OSCILLATOR



A single non-buffered HC inverter can be made to function as a stable oscillator. The crystal may be either a fundamental type (low frequencies) or an overtone model (high frequencies).

Circuit 1 in the diagram shows an oscillator that operates on the third overtone of the crystal. Note that oscillation is possible even on the seventh or ninth harmonic. The 'U' in the type number of the IC indicates an unbuffered output; this type of IC is more suitable for use as an oscillator than the HC model. The crystal has a capacitance, C_{XL}, of 30 pF. This value is important, because the capacitance and inductance L_1 form a resonant circuit that is tuned to a frequency just below the wanted crystal frequency. This arrangement prevents the crystal oscillating spontaneously on its fundamental frequency. The value of resistor R_1 is specified as 3.3 MΩ, but may lie anywhere between 1 MΩ and 10 MΩ.

Circuit 1 may be used with power supplies of 5 V at frequencies up to 40 MHz. The HCU04 chip is not suitable for operation above 40 MHz.

A crystal that oscillates at its fundamental frequency (which may be up to 17 MHz) does not need a tuned circuit. Both circuit 2 and 3 may be used. The difference between the two is the use of either a resistor or a capacitor at the output. For frequencies up to 2 MHz, circuit 3 is more suitable; for higher frequencies (up to 17 MHz), circuit 2 is better.

Calculating the component values for Circuit 1 is simplified by the BASIC program in **Fig. 2**. This program computes the values of C_1, C_2 and C_3 from the input frequency and capacitance, C_{XL}, of the crystal. Normally, the computed alue of C_1 is about half that of C_2 and C_3.

Note that the oscillators shown here are not suitable for use with watch crystals that operate at about 32 kHz.

Design: L. Pijpers
[944013]

```
10 DIM I$(8): PI = 3.141593
20 CLS : PRINT "Calculating crystal oscillators."
30 LOCATE 3, 1: PRINT "An inverter type 74HCU04 (unbuffered) is used."
40 LOCATE 6, 1: INPUT "Crystal frequency in MHz          "; F: IF F = 0 THEN 20
50 LOCATE 9, 1: INPUT "Capacitance C(1) of the crystal (Return=30pF) "; CL
60 LOCATE 12, 1: PRINT "R1 = 3.3  MOhm"
70 IF CL = 30 OR CL = 0 THEN C1 = 27: C2 = 56: GOTO 150
80 IF CL = 8 THEN C1 = 6.8: C2 = 15: GOTO 150
90 IF CL = 12 THEN C1 = 10: C2 = 22: GOTO 150
100 IF CL = 15 THEN C1 = 12: C2 = 27: GOTO 150
110 IF CL = 20 THEN C1 = 18: C2 = 33: GOTO 150
120 IF CL = 50 THEN C1 = 39: C2 = 82: GOTO 150
130 IF CL = 100 THEN C1 = 82: C2 = 180: GOTO 150
140 C1 = .45 * CL: C2 = 2 * C1
150 PRINT "C1 ="; C1; " pF": PRINT "C2 ="; C2; " pF": PRINT "C3 ="; C2; " pF"
160 IF F < 17 THEN PRINT "No need for inductor.": GOTO 180
170 T = 1111 / F: K = T * T / (4 * PI * PI): L = K / C1: PRINT "L1 ="; L; " uH"
180 LOCATE 20, 1: PRINT "More calculations (y/n) "; : INPUT I$
190 IF I$ = "N" OR I$ = "n" THEN END ELSE 20
```

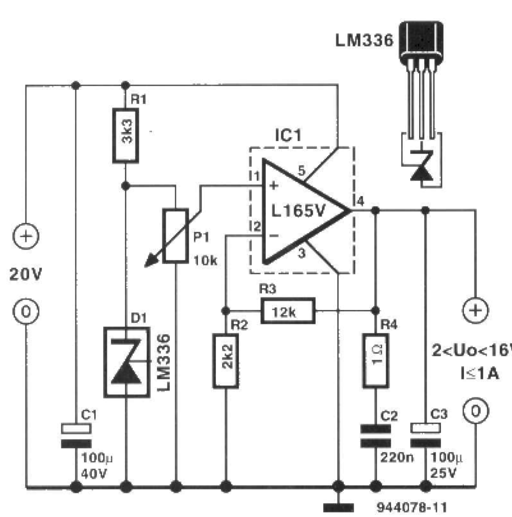
944013-14

NEAR-IDEAL SUPPLY

An ideal power supply maintains its e.m.f., irrespective of whether it is providing current or sinking it. The present supply approaches this ideal. Its output voltage may be set between 2 V and 16 V, while the output current may vary between –1 A and +1 A.

The supply is based on IC_1, a Type L165V operational amplifier from SGS Thomson. A stable reference voltage is provided by zener diode D_1. This voltage can be varied between 0 V and 2.5 V by P_1.

The opamp magnifies its input voltage ×6.45. In theory, its output voltage should, therefore, be 0–16.1 V. In practice, this is not wholly realizable and the output voltage is limited to 2–16 V.



Series network R_4-C_2 prevents the opamp oscillating spontaneously.

The opamp has various internal protection circuits, so that, provided it is mounted on a suitable heat sink of about 4.5 K W^{-1}, nothing serious can go wrong.

Design: H. Bonekamp
[944078]

CENTRONICS INPUT/OUTPUT INTERFACE

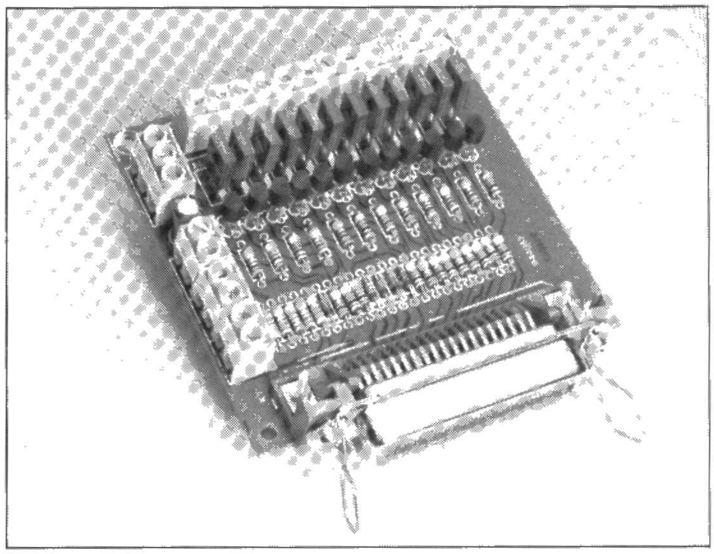
The Centronics interface available on almost any IBM PC or compatible lends itself to simple input/output functions provided the drive capacity of the output lines is considerably boosted. In principle, the five inputs on the Centronics interface can be used straight away because they are TTL-compatible. However, for added security, a resistor and a voltage limiting zener diode may be hooked up as shown here. It should be noted that this works on modern PCs only, i.e., those having CMOS inputs. If you have a very old computer, it is likely to have standard TTL inputs. If so, consider donating it to your nearest science museum, or reduce the value of the protection resistors to a few hundred ohms (or omit them altogether). A simple test to see if this is necessary is to measure the voltage drop across the 2.2-kΩ resistors when applying a '0'. If the voltage drop is greater than 0.8 V, you have an 'old' computer, and changing the resistors to low-value types is in order, as well as driving the inputs from low-impedance sources.

The inputs are pin 11 (BUSY; bit 7), pin 10 (ACK; bit 6), pin 12 (PAPER EMPTY; bit 5), pin 13 (SELECT; bit 4), and pin 15 (not connected; bit 3). The pin numbers refer to the 25-way sub-D connector on the back of the PC. The corresponding pin numbers on the 36-way Centronics connector may be found in the circuit diagram. The logic state of the above bits may be interrogated by reading the LPT base address + 1. In this word, bits 0, 1 and 2 are 'don't care', that is, they do not contain relevant information.

The eight (data) outputs on D25 connector pins 2 through 9 are normally capably of sourcing 2.6 mA or sinking 24 mA. The outputs can be controlled by writing the corresponding eight databits to the base address of the LPT port. In addition to the eight databits, four extra outputs can be addressed via base address + 2: pin 1 (STROBE; bit 0), pin 14 (AUTO FEED; bit 1, pin 16 (INIT; bit 2), and pin 17 (SELECT IN; bit 3). Originally these are control line outputs with an internal pull-up of 4.7 kΩ, and a current sink capacity of 7 mA. Note that the four higher-order bits (4 through 7) must be kept at '0' to prevent an interrupt with unexpected results.

Darlington transistors type BD679 are used to boost the current sink and source capacity of the twelve programmable outputs. The BD679 is capable of switching up to 4 A at a collector voltage of 80 V. In practice, it is recommended to stay below 2 A or so, and fit the transistors with heatsinks if considerable dissipation is expected. Remember that the collector of a darlington can never pull the load lower than 0.6 V, or even 0.8 V at a current between 1 A and 2 A.

The auxiliary voltage used for the BC547 driver transistors is not critical, and may lie between 5 and 15 V. At 5 V, the



total current consumption is about 50 mA.

The loads to be controlled by the Centronics port are connected between the collectors of the BD679s and the positive line of the external power supply (max. 80 V). If inductive loads (such as relay coils) are connected, do not forget to shunt these with (reverse connected) free-wheeling diodes. This is necessary to protect the darlingtons against reverse-emf voltage surges.

As regards programming, you may start in the simplest possible way in BASIC, for instance, with the following program which reads your four inputs, and presents the corresponding value in hexadecimal on the screen:

```
REM read Centronics inputs,
    display in hex
LPT1address=&H378
WHILE 1
cent=INP(LPT1address+1)
PRINT hex$(cent)
WEND
```

Or, equally simple, control the outputs:

```
REM square waves on D0-D7,
    D0 has highest frequency
LPTaddress=&H378
count=0
WHILE=1
OUT LPT1address, count
count=count+1
IF count>255 then count=0
WEND
```

Parts list

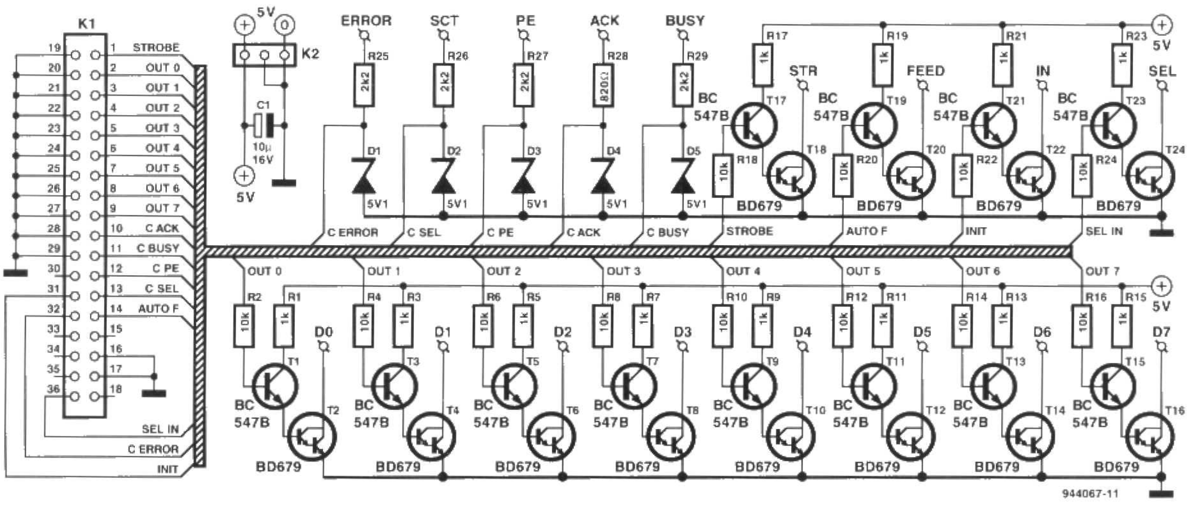
Resistors:

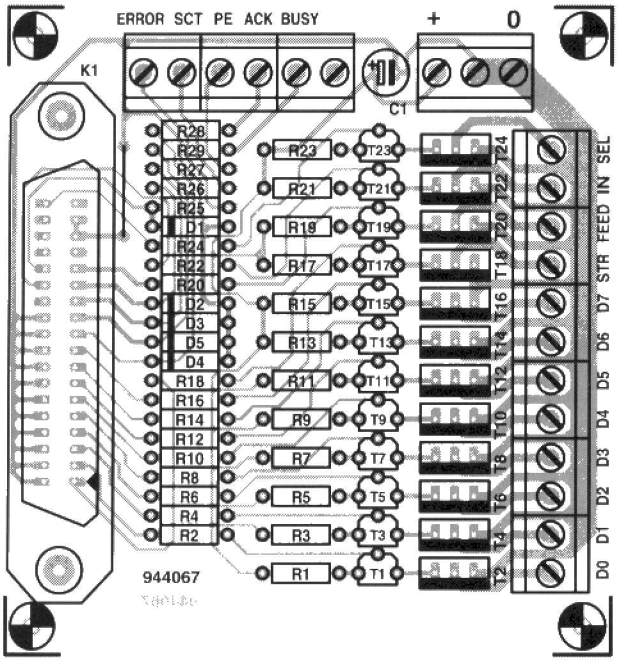
$R_1, R_3, R_5, R_7, R_9, R_{11}, R_{13}, R_{15}, R_{17}, R_{19}, R_{21}, R_{23}$ = 1 kΩ

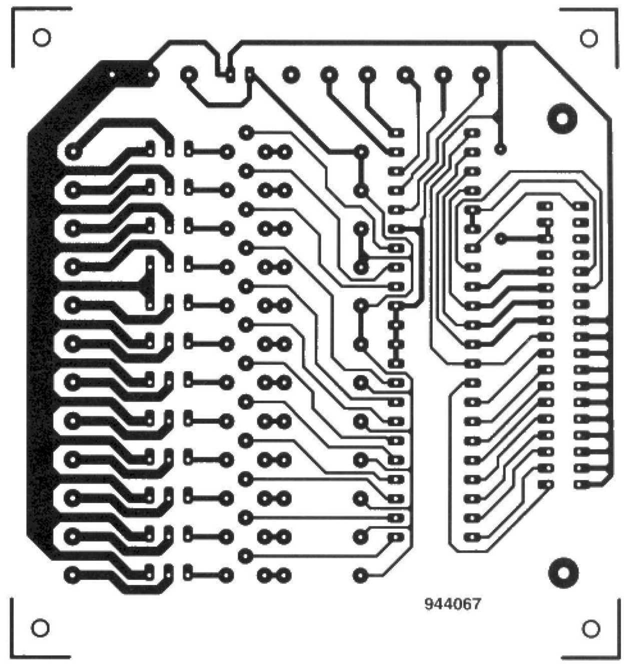
$R_2, R_4, R_6, R_8, R_{10}, R_{12}, R_{14}, R_{16}, R_{18}, R_{20}, R_{22}, R_{24}$ = 10 kΩ

R_{25}–R_{27}, R_{29} = 2.2 kΩ

R_{28} = 820 Ω







Capacitor:
C_1 = 10 µF, 16 V, radial

Semiconductors:
D_1–D_5 = 5.1 V, 400 mW
T_1,T_3,T_5,T_7,T_9,T_{11},T_{13},T_{15},
T_{17},T_{19},T_{21},T_{23} = BC547B
T_2, T_4, T_6, T_8, T_{10}, T_{12},T_{14}, T_{15},
T_{18}, T_{20}, T_{22}, T_{24} = BD679

Miscellaneous:
K_1 = Centronics socket, PCB mount, angled pins.
K_2 = 3-way PCB terminal block, pitch 5 mm.
18 PCB solder pins or 9 PCB terminal blocks, 2-way, pitch 5 mm.

PCB Ref. 944067 (p. 110)

Design: K.M. Walraven)
[944067]

NON-VOLATILE CONTROLLER CHIP

The DS1210 controller chip from Dallas Semiconductor, in combination with a CMOS RAM and a lithium battery, forms a memory module that will retain its data for many years.

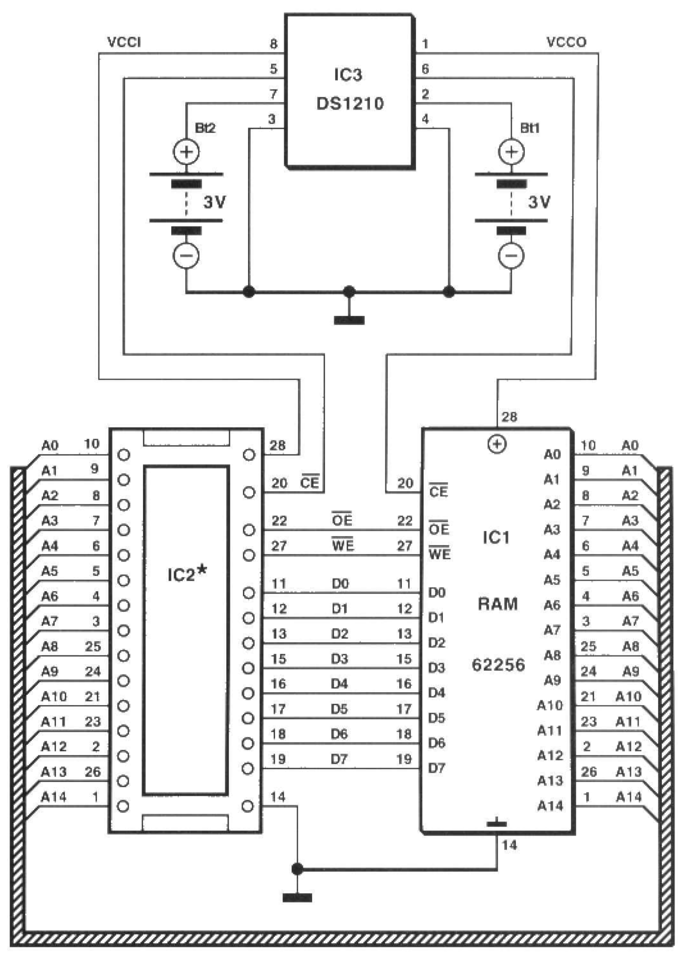
The controller performs several functions. It provides voltage to the RAM IC from V_{cc} or from the battery, depending on the voltage levels. The loss across the internal electronic change-over switch is smaller than 0.3 V.

If the supply voltage fails, the controller prevents any further writing to the memory. The battery is then used as a temporary power supply. To make the circuit absolutely immune to power failure, a second battery may be used, which takes over when the main battery goes flat.

There is a facility to indicate to the user (via the software) the status of the main battery. If that voltage has dropped below 2 V, the controller blocks all read and write operations to the memory from the second read/write cycle onwards. In other words, if it proves impossible to change the content of the memory, the main battery is flat.

The diagram shows how the circuit can be used in an existing apparatus. Note that IC_2 is the socket into which the memory to be protected is to be inserted.

Dallas Semiconductor
Application
[944107]



PEAK A.F. VOLTAGE METER

The meter described can measure peak voltages from a few millivolts to several volts at frequencies up to 200 kHz.

The peak detector is formed by a comparator IC with open-collector output—see **Fig. 1**. Capacitor *C* will be charged to a certain potential. When the level of the incoming signal is below this voltage, the output transistor begins to conduct, whereupon the voltage across *C* will rise rapidly to the level of the negative supply line. When U_C is equal to the input voltage, charging will cease. In this manner, U_C will always correspond to the most negative level of the instantaneous input signal. As *C* is discharged gradually via *R*, U_C will follow any variations in the input signal. Of course, the circuit in **Fig. 1** can measure only half the peak-to-peak voltage. Adding an identical circuit for the positive halves enables the whole signal to be measured.

In **Fig. 2**, IC_1 is the positive-peak detector and IC_2 the negative-peak detector, whose 'memories' are C_1 and C_2 respectively. Low-pass filters R_3-C_3 and R_6-C_4 remove any unwanted peaks. The positive value is inverted by IC_3 and summed with the negative half in IC_4, where the signal is inverted again.

The top end of the frequency range is limited only by the speed of IC_1 and IC_2; in the prototype, the upper frequency was about 200 kHz, even with rectangular and triangular signals.

The lower end of the frequency range is determined by the values of C_1 and C_2. With the values as shown, the lower limit was about 500 Hz. If measurements down to 20 Hz are desired, the value of these capacitors must be increased to 220 μF—but this will be at the cost of the reaction time. Following a rise in the input signal of 1 V will then take 2–3 seconds.

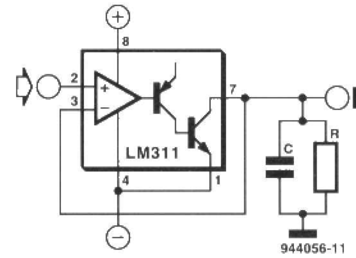
Note also that capacitors C_1 and C_2 are bipolar types. Only if there is certainty that the input signal will always be free of direct voltage can normal electrolytic types be used.

The supply voltage may be as low as ±6 V, but if input signals exceeding 4 V_{pp} are to be measured, or if the signal has an appreciable d.c. offset, it is advisable to use a higher supply voltage. Although not strictly necessary, it is also advisable to stabilize the supply lines. Since the current drain does not exceed 20 mA, two 9 V batteries will suffice.

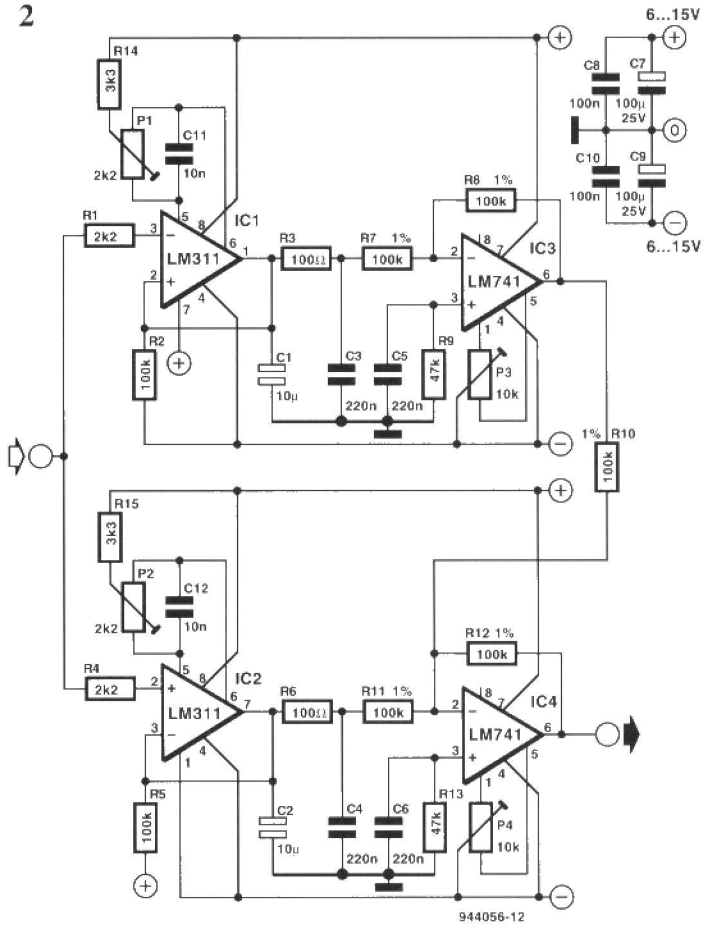
To calibrate the meter, short-circuit the input to ground, connect a millivoltmeter across C_1 and adjust P_1 for a reading of 0 V. Next, do the same with the voltmeter across C_2 and adjust P_2. This needs to be done carefully, because it is a matter of just a few millivolts. Then, adjust P_3 and P_4 for a voltage of 0 V at pin 6 of IC_3 and IC_4 respectively. Remove the short-circuit from the input.

Design: V. Mitrovic [944056]

1



2



COMPRESSOR FOR GUITARS

Many guitarists still swear by the good old valve amplifier. It is fact that valve amplifiers produce a sound different from that of a solid-state amplifier. This is caused mainly by the fact that in valve amplifiers the transition from the linear to the non-linear part of the operating range is more gradual than in solid-state amplifiers, in which the transition is sudden. In other words, in valve amplifiers, the distortion increases with increased drive levels and may thus be seen as a sort of dynamic compression. This phenomenon does not occur in solid-state devices, where the transition is quite sudden: the resulting sound is unpleasant.

Before the arrival of voltage-controlled IC amplifiers, dynamic compressors were often designed with diodes as the control element. Since a certain degree of distortion is desirable, diodes are very suitable for use in a guitar compressor with 'valve sound'.

In the diagram, input amplifier IC_1 drives two pairs of diodes via R_5: D_1-D_3 for the positive half periods and D_2-D_4 for the negative half periods. Capacitors C_3 and C_4 short-circuit any a.c. signals: they hold the level of the control voltage constant and thus determine the speed of the control. The diodes obtain a bias voltage (which is independent of the drive level) from the output of IC_{2a} via D_5 and D_6.

When S_{2a} is open, the dynamic compression is disabled.

The stage based on IC_{2b} compensates the output level at

various positions of the switches and potentiometers (the two halves of which turn in opposite directions). It is thus possible to take the output signal directly from the output of IC_{2a}.

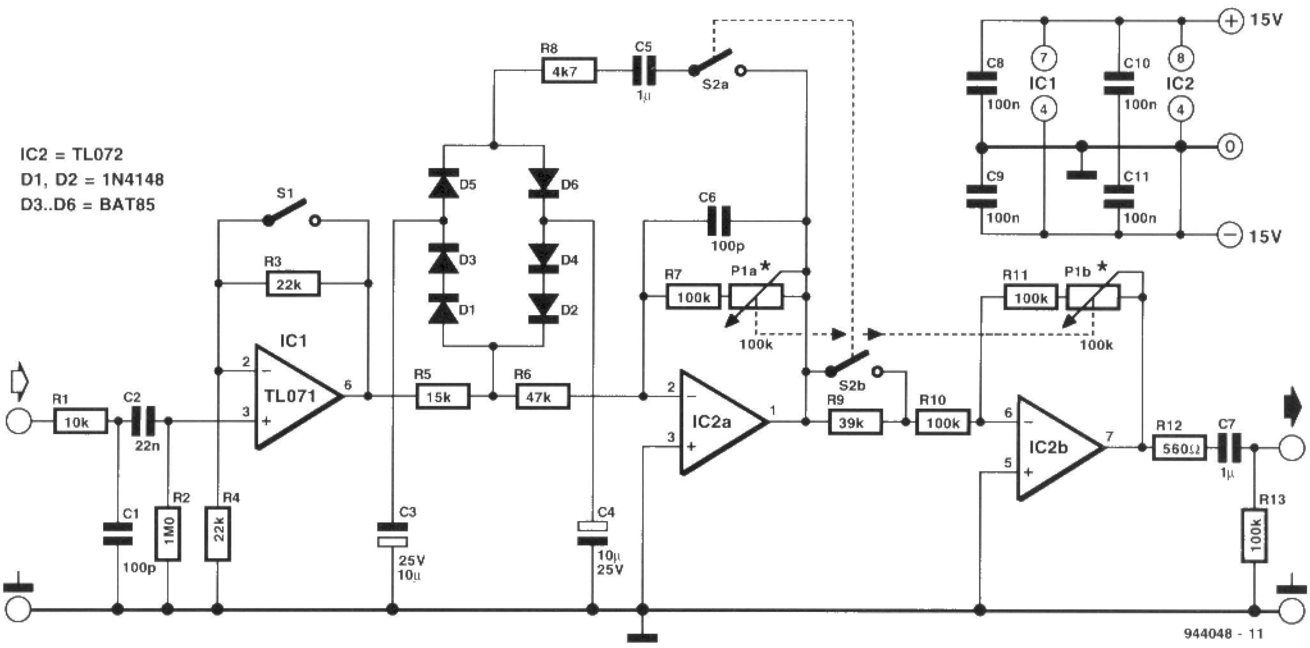
The onset of compression is set with P_1. This control should be set, by trial and error, to a position where the 'plucking peaks' are just beginning to distort.

The amplification of IC_1 is altered with S_1 to enable instruments with a single as well as a double coil to be used.

Although the prototype performed to the taste of the designer, individual modifications are, of course, possible. Note, however, that the BAT85 diodes must not be replaced by common-or-garden diodes like the 1N4148, which produce a far less pronounced 'valve effect'. This is because the distortion sets in more abruptly and at a higher level.

Design: W. Teder

[944048]

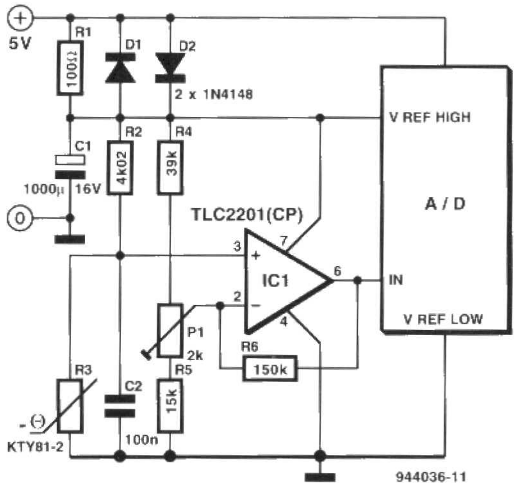


SENSOR INTERFACE

Connecting a sensor to an A-D (analogue-to-digital) converter is not always as straightforward as may be expected. Often, an interface is required to amplify the signal and to provide some offset control. The present circuit was designed on that basis.

Most A-D converters have a reference input; the voltage applied to this is fed internally to a potential divider. This means that the converter does not work with absolute values, but with the ratio of input voltage to reference potential.

If, for instance, a temperature sensor is required, it is best to base this on a resistor with positive temperature coefficient (PTC) connected in a bridge circuit. If the supply to the bridge is taken from the same source as the reference voltage, the reading of the A-D converter will be immune to small variations in the level of the reference voltage. In other words, the reference voltage need not be regulated.



If a sensor sensitivity of about 0.75% °C^{-1} is sufficient, a bridge as described earlier will be adequate. If greater precision is required, the bridge voltage must be amplified, which can be arranged with any opamp that can operate from +5 V. In the prototype, a modern, low-dissipation type was used as shown in the diagram. This model, which has very low noise and very small offset, can be driven to 0.2 V with small loads (1 mA) from a negative as well as a positive supply.

The A-D converter used an unregulated +5 V supply. Diodes D_1 and D_2 make sure that the reference voltage remains close to the supply voltage.

One branch of the bridge contains preset potentiometer P_1, while the other contains the PTC resistor, R_3. Resistor R_2 does not only form part of the bridge, but also serves to make the temperature-to-voltage characteristic of R_3 linear. With the type of PTC resistor in the diagram, R_2 needs to be 4.1 kΩ to obtain a linear curve around 20 °C. If the KTY81-1 is used, the value of R_2 should be 2 kΩ. Capacitor C_2 suppresses any noise.

The amplification of the opamp is determined by feedback resistor R_6, whose value is up to individual requirements (within reason). With values as specified, the amplification is about ×14.

The circuit draws a current of only a few mA.

Design: K. Walraven

[944036]

LEAD-ACID BATTERY CHARGER

Apart from use as a standard charger, the present unit may also be used for continuous trickle charging to keep a 12 V lead-acid battery in top condition.

The charger is based on a precision voltage source, which has been given a negative temperature coefficient by a temperature sensor. This means that the charging voltage is reduced when the ambient or battery temperature rises.

According to the electrical firm of Bosch, a temperature coefficient of –8 mV °C^{-1} is the most beneficial for charging a lead-acid battery. In the present circuit, this is achieved by using a common-or-garden transistor as sensor.

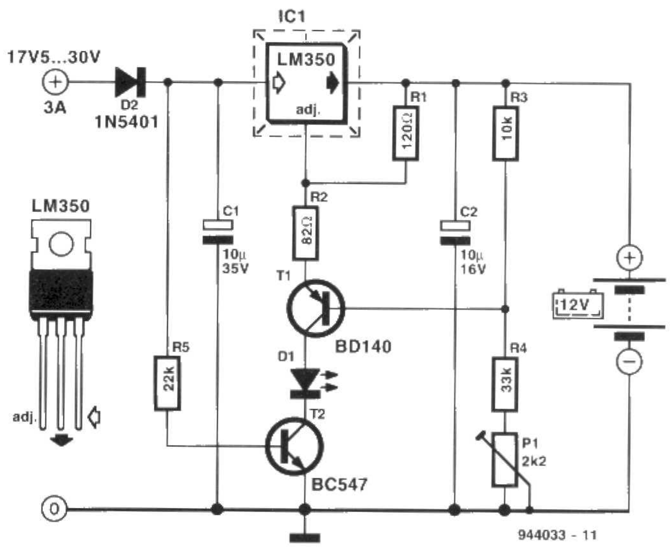
The operation of IC_1 depends on the property that three-pin regulators tend to keep the voltage difference between their input and output at a constant 1.25 V. This means that the current through R_1 is constant. Normally, this characteristic is made use of to set the desired output voltage by means of a fixed resistor between the adj(ust) pin and ground. In the present circuit, this resistor is made variable by inserting T_1 in the link. To keep the circuit stable, potential divider, R_3-R_4-P_1 controls the base voltage of T_1.

Since the base-emitter junction of T_1, in common with every other semiconductor, has a temperature coefficient of about –2 mV °C^{-1}, the output also has a negative temperature coefficient. This is, however, four times as large because the base-emitter variation of T_1 is multiplied by the scaling factor of R_3-R_4-P_1.

Diode D_1 shows whether the power is available.

Transistor T_2 prevents the battery being discharged via R_1 in the absence of a supply voltage (T_2 is switched off).

The wanted output voltage is set between 13.5 V and 14.5 V with P_1. This range may be shifted to some extent by altering the value of R_4.



To prevent T_1 being warmed by its own base current, it is advisable to fit the transistor on a small sheet of metal.

If T_1 is intended to monitor the ambient temperature only, it suffices to mount it just in the open air. If it is intended to compensate a rise in battery temperature, it must be mounted as close to the battery as possible.

Linear Technology Application

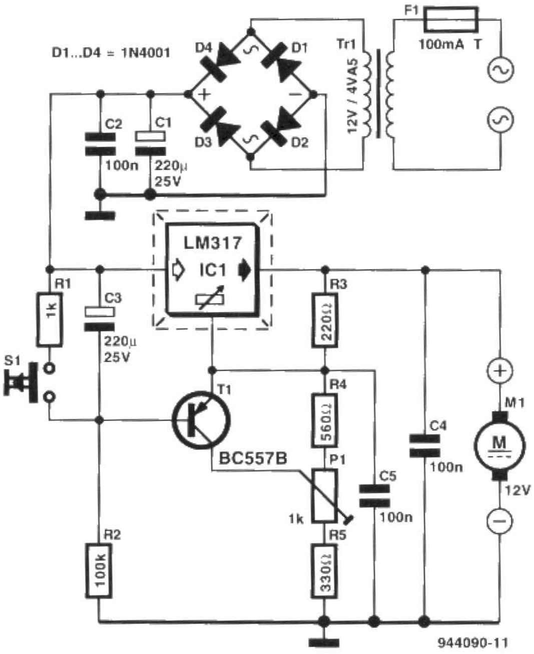
[944033]

SOLDER VAPOUR EXTRACTOR

Soldering is an unhealthy activity: when the tin melts it releases harmful vapours. It is, therefore, advisable, to install a small 12 V d.c. extractor fan on or near the workbench. The present circuit is intended to control the fan.

The rotational speed of the fan is adjusted to a few hundred r.p.m. with P_1. Pressing push-button switch S_1 during soldering raises that speed to maximum to give added extraction of the vapours. The length of time during which the fan rotates at maximum speed depends on the time constant R_2-C_3. The fan will also rotate at maximum speed for a time R_2-C_3 when it is switched on. After that, its speed drops to a few hundred r.p.m.

The circuit is a typical application of an LM317. The output voltage is determined by P_1 when T_1 conducts and by potential divider R_3/(R_4-P_1-R_5) when the transistor is off. With values as specified, the output voltage ranges from about 4 V to around 11 V.



When S_1 is pressed, the potential across C_3 drops to 0, whereupon T_1 stops conducting. After a little while, the voltage across C_3 has risen sufficiently to cause T_1 to begin to conduct again. It then forms a resistance in parallel with R_4 and the upper half of P_1. As T_1 begins to conduct harder and harder, its parallel resistance becomes smaller and smaller until it is a short-circuit. The output voltage then depends on the ratio between R_3 and the lower half of P_1 plus R_5. The potentiometer is thus used to set the minimum output voltage. With values as specified, the minimum output voltage can be set between 4 V and 8 V. This minimum voltage is used to drive the fan at low speed. If the fan stops, adjust P_1 to give a slightly higher output voltage.

Design: A. Rietjens

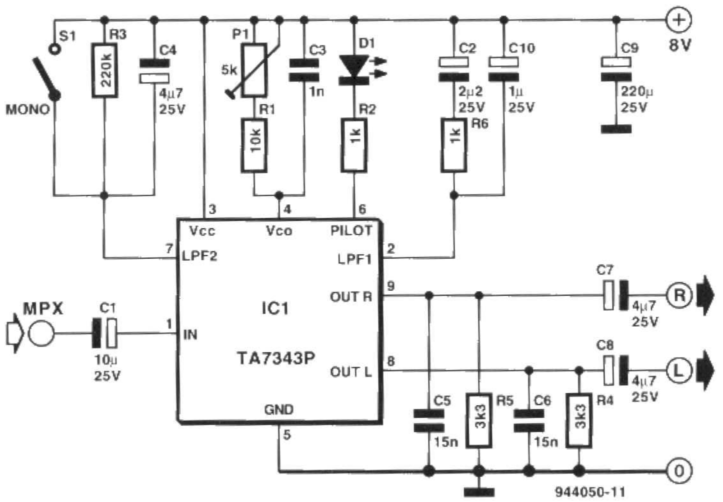
[944090]

STEREO DECODER

Extracting the separate left-hand and right-hand signals from a multiplexed stereo signal is a complex process. It requires an auxiliary carrier of exactly the same frequency (38 kHz) and phase as that in the transmitter. A voltage-controlled oscillator (VCO) by itself is not good enough. It needs to be synchronized with the 19 kHz pilot tone in the multiplexed (MPX) signal. Furthermore, a mono/stereo converter controlled by the pilot tone, preferably with an optical indicator, is required.

Fortunately, ready-made IC stereo decoders have been available for some time. Nevertheless, these decoders normally require quite a few external components and, moreover, their calibration is often not simple.

Modern decoder ICs, however, are simplicity itself. As the diagram shows, one of these, Toshiba's TA7343P, needs only a few external components.



The MPX signal is applied to pin 1, whereupon the separate left-hand and right-hand signals are available at pins 8 and 9 respectively. When the internal pilot-tone detector registers the presence of the 19 kHz signal, D_1 is driven into conduction and the internal 'stereo switch' is actuated. When there is no pilot tone, D_1 remains off and the circuit remains in the mono mode.

Manual switching to the mono mode is possible with S_1. When this switch is closed (pin 7 to +8 V), both outputs carry the L+R signal.

The VCO is adjusted with P_1 and does not require any instruments or expertise: simply tune the receiver to a stereo transmitter and turn P_1 till D_1 lights.

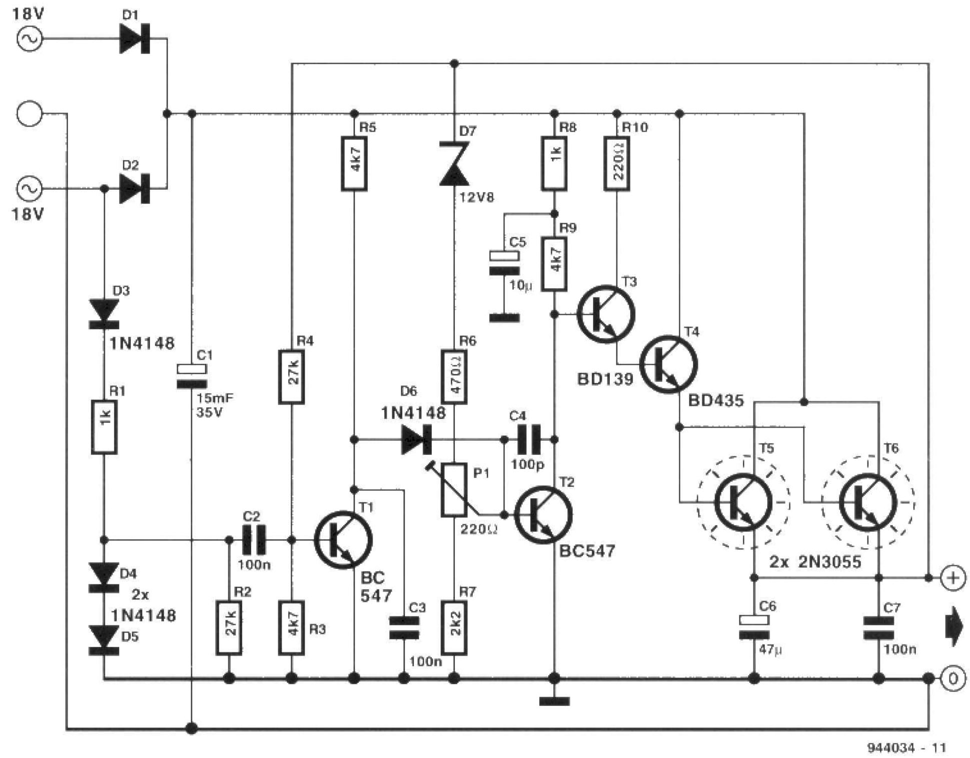
In spite of its simplicity of application, the parameters of the TA7343P are very good. Harmonic distortion is 0.08% (typical); auxiliary carrier suppression is 70 dB; and the signal-to-noise ratio is 74 dB. These figures were measured with a supply voltage of 8 V.

The supply voltage can range from 3.5 V to 12 V. The current drain (excluding that through the LED) is 11–18 mA, depending on the supply voltage.

Toshiba Application

[944050]

DISCRETE POWER SUPPLY



In these days of integrated voltage regulators, many constructors will enjoy this old-fashioned power supply built from discrete components. It provides an accurately presettable output voltage of 13.8 V at a peak current of 10 A and is short-circuit-proof.

Ignoring the stage round T_1 for a moment, the design is traditional. Transistor T_2, in conjunction with potential divider R_6-R_7-P_1 and zener diode D_7, forms a variable voltage source that is followed by three emitter followers. The last of these, T_5 and T_6, consists of two parallel-connected golden oldies Type 2N3055, which together provide the output current.

What is new is the manner of current limiting provided by T_1. The base voltage of this transistor is derived from the output voltage via R_4. If the output voltage drops below 5 V, for instance through overload, T_1 will switch off. Transistor T_2 is then driven hard via D_6, which results in T_3–T_6 only justconducting, so that the output voltage and current become negligibly small.

To get out of this situation, a 50 Hz signal is applied to the base of T_1 via D_3. The level of this signal is limited to 1.3 V by D_4 and D_5. This causes T_1 to switch on periodically, which will reenable the circuit if the overload or short-circuit at the output has been removed. This means, of course, that even during a short-circuit fairly large current pulses flow through the load, but these are so short (about 2 ms) that they can not cause any harm.

Instead of the usual emitter resistors, the connecting wires to the emitters of T_5 and T_6 function as potential divider. They must, therefore, both be about 10 cm long and have a diameter of 0.7 mm.

Because of the fold-back current limiting, the heatsink for T_5 and T_6 need not be large: a 10 cm long Type SK01 ($2.5\ \mathrm{K\,W^{-1}}$) will suffice.

Choose transistors with a T03 case for T_5 and T_6 and mount with the aid of insulating washers.

Preset P_1 should preferably be a multiturn type.

Buffer capacitor C_1 may consist of three parallel-connected 4700 µF electrolytic capacitors: this is at least as good as a 15 mF type and much cheaper.

Diodes D_1 and D_2 must be 10 A types: it may be cheaper to use two diodes from a 10 A bridge rectifier.

Finally, make sure that the large currents can not enter the control circuits.

Design: Altai
[944034]

THREE-PHASE INDICATOR

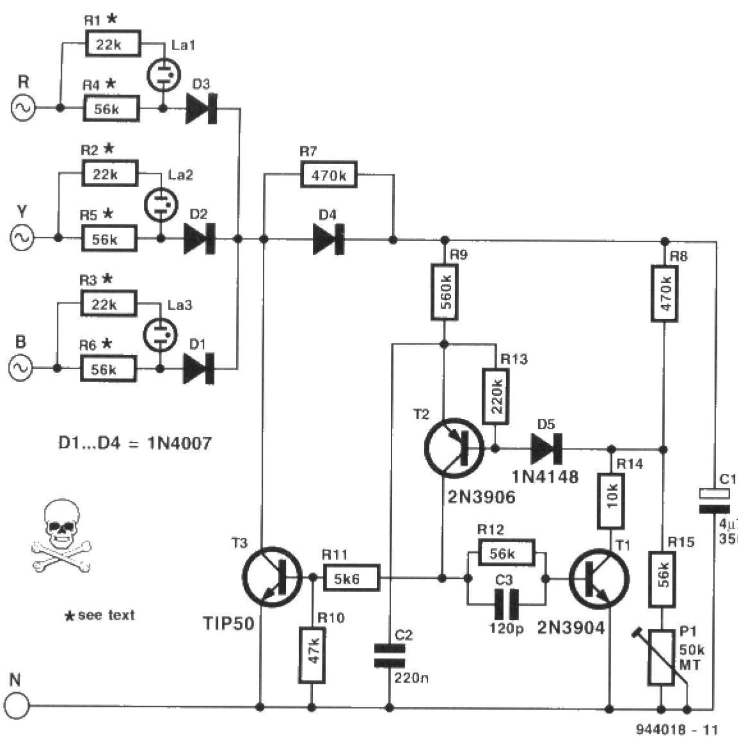
When a three-phase mo-tor is being installed, it is essential to know the sequence of phases R, Y and B, since an error may have disastrous consequences. The letters R, Y and B are abbreviations of 'red', 'yellow' and 'blue', that is, the colours used to identify the three phases. Also, red-yellow-blue is the sequence universally adopted to denote that the e.m.f. in the yellow phase lags that in the red phase by 120° (a third of a cycle), and the e.m.f. in the blue phase lags that in the yellow phase also by 120°.

In the present circuit the phases are indicated by three neon lights mounted in a circle. If this 'running light' moves clockwise, the phase sequence is correct; if it moves anti-clockwise, the sequence is incorrect.

Each neon lamp is in series with a phase via a resistor. The three phases are simply interconnected after being rectified by D_1–D_3. Their junction is connected on and off to N(eutral) by T_3. The remainder of the circuit arranges for the switching to occur in a manner by which the neon lamps are short-circuited to neutral in the correct sequence.

Transistor T_3 is controlled by a monostable formed by T_1 and T_2, which provides a switching signal at a frequency of about 48 MHz. It works as follows. After the phases have been rectified by D_1–D_4, capacitor C_2 is charged via R_9. As soon as the potential across C_2, and thus that at the emitter of T_2, rises above the level set by R_8-R_{15}-P_1, T_2 begins to conduct. This causes C_2 to be discharged via T_2, R_{11} and the base-emitter junction of T_3. Transistor T_3 is then on briefly so that one of the neon lamps, depending on which phase is active at that instant, is short-circuited to neutral

and lights. The lamp lights until C_2 is virtually discharged. This is because the collector of T_2 keeps T_1 switched on as well, so that junction R_8-R_{15} is also at neutral. When C_2 has been discharged completely, T_2 and T_1 switch off and the neon lamp goes out. This continues until the next phase becomes active.



Provided that good-quality components are used for R_8, R_9, R_{15}, P_1 and C_2, the monostable will ensure a reliable indication.

The circuit is calibrated as follows. Connect it to the three-phase supply (neutral first!) after having turned P_1 fully aticlockwise. Then, slowly turn the preset clockwise until the running light formed by the neons appears to be at a standstill. Next, turn P_1 a little further clockwise until the neon lights appear to move in a clockwise direction at about two revolutions per second. Lock the preset in that position with some lacquer.

Because of the required accuracy, R_8, R_9 and R_{15} must be metal film types; C_2 should be a polyester type and P_1 must be a multiturn model. The neon lamps should have their series resistors integrated. Capacitor C_1 must be a high-voltage type. Resis-tors R_4–R_6 must be special 500 V types (it is, of course, possible to connect two standard 250 V types in series).

Finally, the entire circuit is connected directly to the three-phase supply so the utmost caution is necessary during construction, testing and calibration. It must be installed in a good-quality synthetic enclosure.

Design: R. Kähne
[944018]

EXPERIMENTATION BOARDS FOR PICs

In view of the popularity of general purpose PICs (Peripheral Interface Controllers), two experimentation boards are presented: one for the 18-pin models (**Fig. 1a**), and the other for the 28-pin models (**Fig. 1b**).

The layout of both boards is given in **Fig. 2**; the board available from our Readers' Service contains both versions. The boards are indispensable for the practical application of the new PIC Programming course started in this issue.

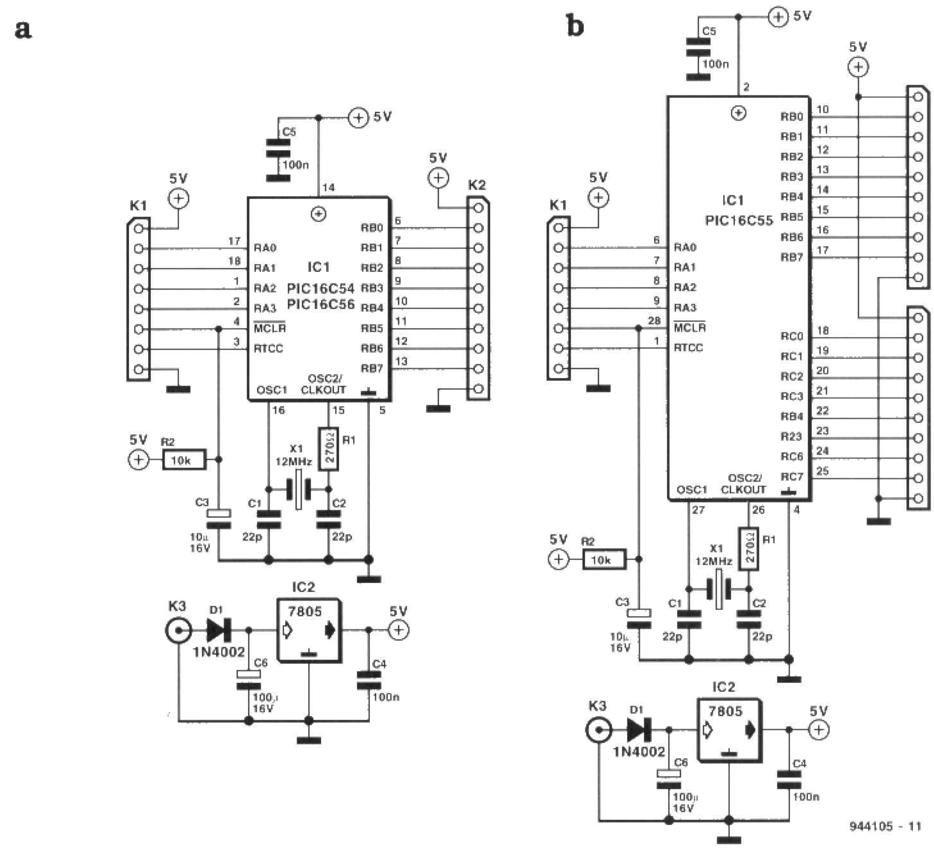
Each board has provision for the 5 V power supply with polarity reversal protection: any mains adaptor that can provide 9–15 V and a current of 300 mA is suitable.

The board also has provision for the 8 MHz clock generator, reset circuit and decoupling capacitor.

The pin headers on the board ensure that access to the I/O lines is good.

Finally, extra soldering pads are provided for any additional hardware that may be needed.

Note that in the EPROM version of the PICs no choice has been made as regards the type of oscillator. This means that during the configuration of the fuses (in the programming) the correct type of oscillator will have to be chosen and programmed (here, the XT).



Parts list

Resistors:
R_1 = 270 Ω
R_2 = 10 kΩ

Capacitors:
C_1, C_2 = 22 pF
C_3 = 10 µF, 16 V
C_4, C_5 = 100 nF
C_6 = 100 µF, 16 V

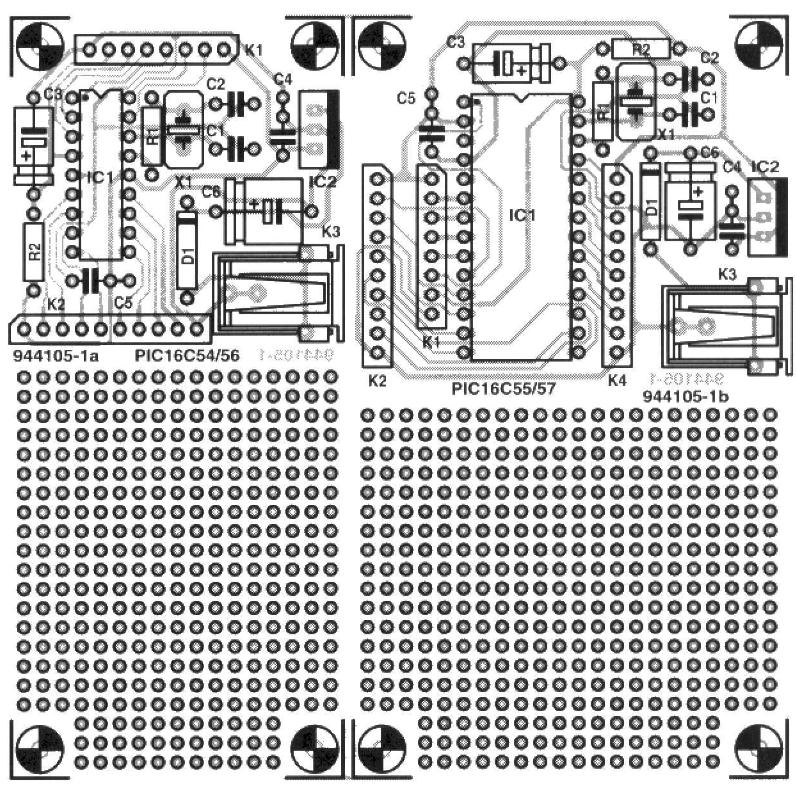
Semiconductors:
D_1 = 1N4002

Integrated circuits:
IC_{1a} = PIC16C54/56 (board a)
IC_{1b} = PIC16C55/57 (board b)
IC_2 = 7805

Miscellaneous:
K_1 = 8-way single row header
K_2 = 10-way single row header
K_3 = mains adaptor connector
K_4 = 10-way single row header

(board **b** only)
X_1 = crystal, 8 MHz
PCB Ref. 944105 (p. 110)

Design: A. Rietjens
[944105]



SWEEP GENERATOR

Phase locked loop IC Type CD4046 enables a simple rectangular wave generator with variable frequency to be designed as shown in the diagram. Careful computing and choosing of the component values results in a usable sweep generator.

Capacitor C_1 is charged via R_1; the resulting potential across it is applied to linked pins 9 and 14 of the IC. Pin 14 is the input to the phase comparator, while pin 9 gives access to the control input of the voltage-controlled oscillator (VCO).

When the input to the IC becomes logic high, the bistable in the phase comparator is reset. This results in the level at pin 13 going high and C_1 being discharged via T_1.

At the same time, the high level at pin 13 causes C_2 to be charged via R_3. When C_2 has attained a given charge, pin 3 goes high, so that the bistable is reset.

Diode D_1 then causes C_2 to be discharged rapidly. At the same time, C_1 is charged again and the whole cycle repeats itself. The potential across C_1, which has an exponential waveform, ultimately provides the sweep of the VCO frequency.

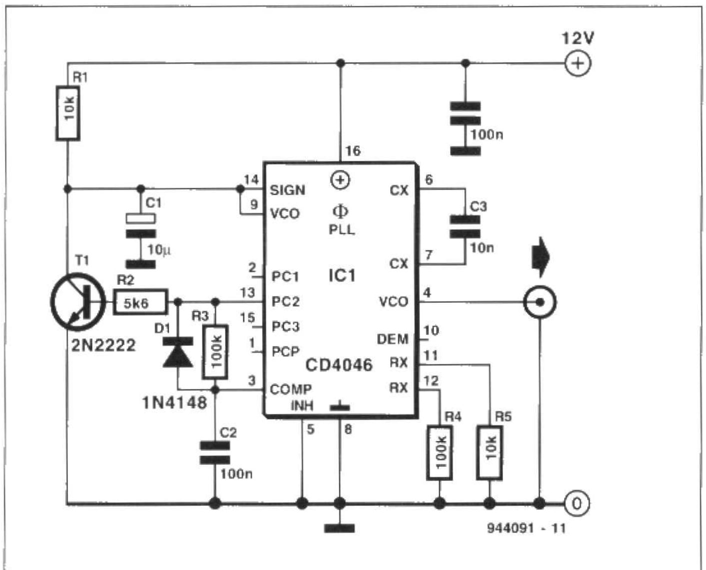
The minimum VCO frequency depends on the values of C_3, R_4 and R_5; with values as specified, it is 2.2 kHz.

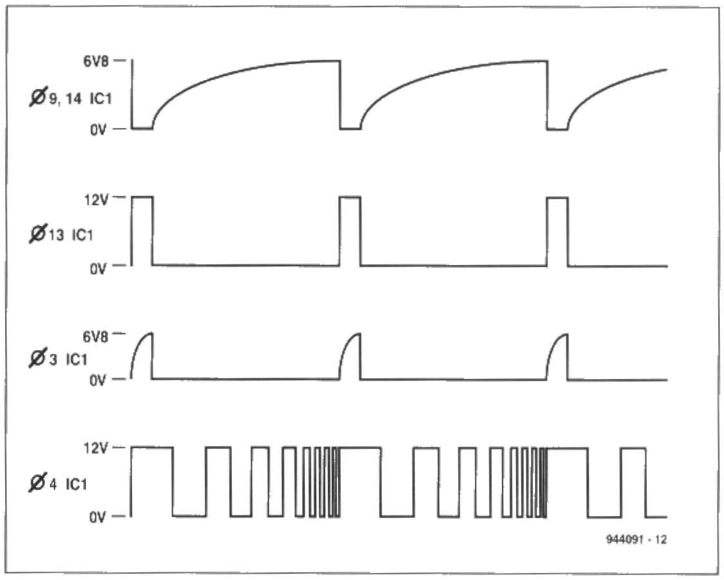
The maximum frequency depends on C_3 and R_4; with values specified it is 11 kHz.

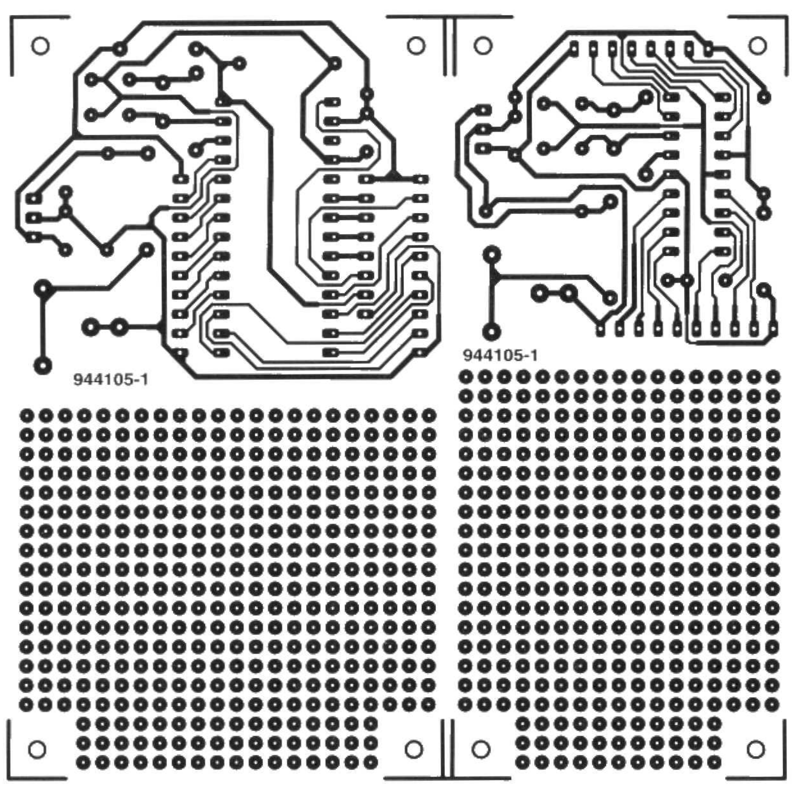
The sweep time of about 600 ms (with a supply voltage of 12 V), is determined by C_1 and R_1.

The supply voltage may be 7–15 V. The circuit draws a current of not more than 5 mA.

Design: M. Nagaraj
[944091]







POLARITY REVERSAL PROTECTION

Reversal of the supply lines to a circuit normally has disastrous consequences. The protection described here can prevent such a disaster occurring. It consists of an enrichment MOSFET that is connected close to the negative supply input. The transistor conducts only when it receives a positive gate voltage from the positive supply line via R_1.

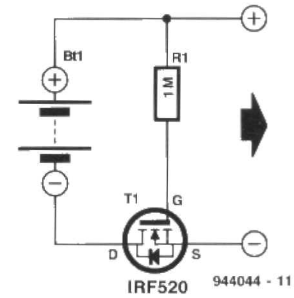
Use is made of an n-channel transistor, which usually has a lower $R_{ds(on)}$ than p-channel types. This low transfer resistance is vital because it means that even with large loads the power loss is kept low.

If breaking into the negative supply line is difficult, a p-channel device can be used in the positive supply line.

There are several n-channel FETs that are suitable for the present application; the IRF520 is but one of them – the IRF540 and BUZ11 are two alternatives.

Designing the protection so that, strictly speaking, the current flows in the wrong direction (from source to drain) is deliberate, since it prevents the internal protection diode conducting when the polarity of the supply is reversed accidentally. This diode will, however, conduct when the current becomes large. This is not much of a problem, because it can cope with the same level of current as the drain-source channel. In the IRF520, this is 9.2 A; the IRF540 can handle up to 28 A. The table shows the potential drop with and without heat sink: it is clear from this that at relatively large currents a heat sink (21 $K W^{-1}$) is a sensible addition. The figures in the table were obtained with a battery voltage of 12 V; if a lower voltage is used, the drops across the FET may be slightly higher.

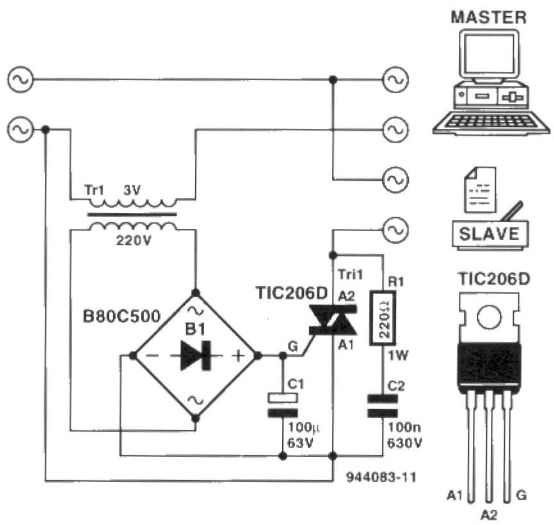
Design: T. Giesberts
[944044]



Current (A)	Drop without heatsink (mV)	Drop with heatsink (mV)
0.012	2.2	2.4
0.120	21.9	26
0.240	43.3	50
0.450	87.3	90
0.980	223	210
1.870	500	420
2.610	1 100	...
2.740	...	640

SLAVE SWITCH

It is often handy if two or more sets of equipment that work together, such as a computer and a printer, or a tuner, amplifier and CD player are switched on and off together. This can be done with the aid of a slave switch, which is easy to build. The simplest type uses a resistor in series with the mains lead as sensor. As soon as a current flows, a voltage is developed across the resistor and this is used to switch on a triac. This arrangement is not only wasteful (power dissipation in the resistor), but also dangerous in that there is a risk of the entire load current flowing through the gate of the triac, which definitely can not cope with this.



It is much better to use a current transformer as sensor. In the present circuit, an old 1 A bell transformer with secondary voltages of 3 V, 5 V and 8 V was used. The 3 V winding is connected in series with the (neutral) mains line to the computer (amplifier). Note, however, that with certain computer-printer combinations, the printer should, according to the manufacturer, be switched on first. As soon as the computer is switched on, the secondary current will be transferred to the primary winding. Since this winding has many more turns than the secondary, the current through it is much smaller than that in the secondary. For a voltage ratio of 3:240, the transformer ratio is 1:80, that is, the primary current is $^1/_{80}$ of the secondary current. This current is large enough, however, to start a sensitive triac like the TIC206D.

The starting current of the triac has the same waveform as that through the primary of the transformer. If this is a clean sine wave, the triac will strike just after the zero crossing as it should. In other cases, the triac might not strike until much later and that is not the idea. To ensure that the triac conducts continuously, a bridge rectifier and reservoir capacitor have been added.

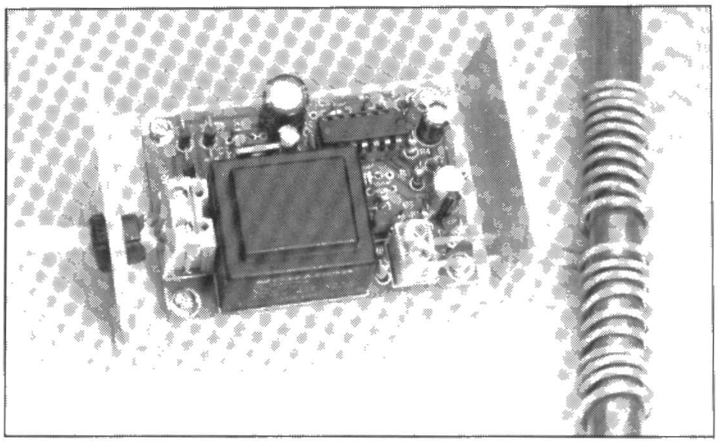
The prototype needed a sinusoidal current of about 100 mA to start the triac. The voltage loss amounted to 95 mV, which increased to 295 mV with a current of 500 mA. Without a transformer, that is, with a series resistor, this would have been at least 2 V. Larger currents may be obtained by the use of the 5 V or 8 V winding. Note, however, that the current must not exceed the rating of the transformer.

Since mains voltage is present in the circuit, great care must be taken when working on the circuit, which, when finished, must be enclosed in a non-metallic housing.

There is a further danger in this circuit in that the transformer is used 'the wrong way around', so that when it is not loaded, very high voltages may arise across the primary winding. It is, therefore, strongly advisable to use a bell transformer, because this complies with stringent safety requirements.

Design: K.M. Walraven
[944083]

WATER SOFTENER



Exposing hard water to an electrical field causes calcium-carbonate crystals and other minerals to join and form larger crystals. At least, that is what researchers claim to have discovered in their efforts to combat the cause and ill effects of hard water.

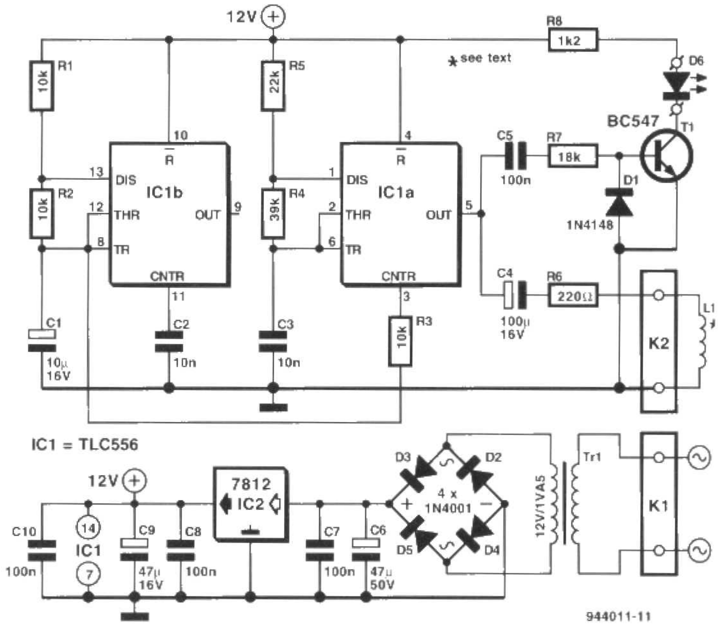
The good news is that the risk of calcium compound deposits reduces with crystal size. If your home is on a hard water supply (which depends on the area you live in), you will probably be aware of the disastrous effects of solidified calcium compounds, which form green deposits on just about anything that runs hot and supplies water, i.e., heating elements in washing machines (expensive to replace!), taps, shower heads and kettles. Water softeners are available commercially, but unfortunately cost an arm and a leg.

A dual astable multivibrator (AMV) type TLC556 is used in the softener circuit to generate a swept-frequency rectangular waveform which is applied to a coil, L_1, wound around a length of plastic water tubing where the water supply enters your home. Alternatively, the coil is fitted on the water inlet tube of a certain apparatus, for instance, a washing machine that is to be protected against hard water. The oscillator output pulses are rectified by D_1, and cause D_6 (a green LED) to light as an assurance that the coil is fed with pulses. The oscillator frequency is swept between about 800 Hz and 2.5 kHz. The coil, L_1, consists of 14 turns of solid, insulated wire, spaced at about the diameter of the water pipe. Some experimentation with different types of electromagnetic field that work on the hard water is worthwhile.

Figure A shows two open-ended coils connected to K_2. This is the traditional arrangement of forming an electromagnetic field, i.e., a field with an electric and a magnetic component. **Figure B** shows three coils connected to K_2. This causes two mutually interacting electromagnetic fields. Lastly the arrangement in **Fig. C** produces a magnetic field only. All of these arrangements should be tried to ascertain which one gives the best results on plastic or copper water pipes. All coils are 14 turns each, spaced as stated above.

The circuit is powered by a conventional mains supply consisting of a transformer, a rectifier, and a 12-V voltage regulator.

Since the mains voltage is present on the printed circuit board (at the primary pins of transformer Tr_1), the completed PCB should be fitted into a robust plastic case. The coil wires are kept as short as possible and connected to PCB terminal block K_2. The circuit can safely be assumed to work if the green LED lights.



944011-11

Parts list

Resistors:
R_1, R_2, R_3 = 10 kΩ
R_4 = 39 kΩ
R_5 = 22 kΩ
R_6 = 220 Ω
R_7 = 18 kΩ
R_8 = 1.2 kΩ

Capacitors:
C_1 = 10 µF, 16 V, radial
C_2, C_3 = 10 nF
C_4 = 100 µF, 16V, radial
C_5, C_7, C_8, C_{10} = 100 nF
C_6 = 47 µF, 50 V, radial
C_9 = 47 µF, 16 V, radial

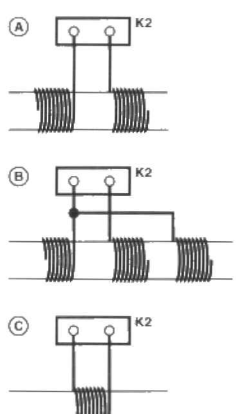
Inductors:
L_1 = 14 turns; see text

Semiconductors:
D_1 = 1N4148
D_2–D_5 = 1N4001
D_6 = green LED
T_1 = BC547

Integrated circuits:
IC_1 = TLC556
IC_2 = 7812

Miscellaneous:
K_1 = two-way PCB terminal block, pitch 7.5mm.
K_2 = two-way PCB terminal block, pitch 5mm.
Tr_1 = 12V 1VA5 mains transformer, e.g., VTR1112 (Monacor/Monarch).
PCB Ref. 944011 (p. 110).
Enclosure: size approx. 37x57x90mm.

Design: L. Pijpers
[944011]



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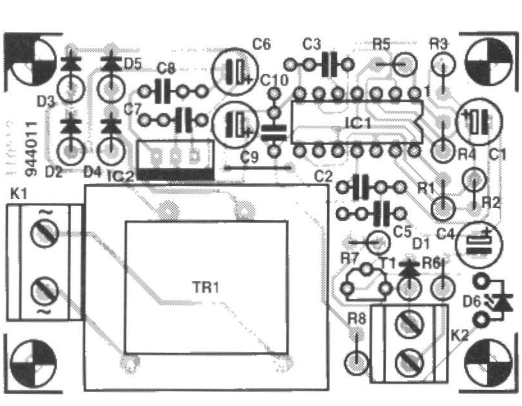
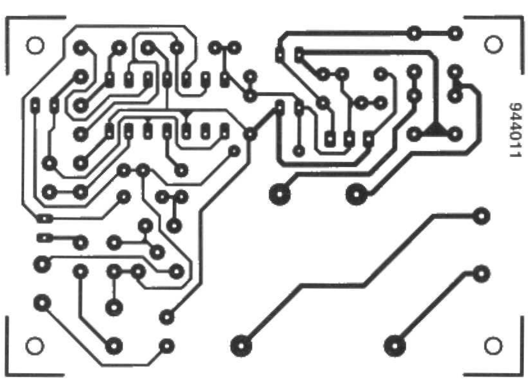




CAR BATTERY MONITOR

The lead-acid battery forms the heart of the electrical system in any car. If you are on a camping or caravanning holiday, it can happen that the battery is discharged so deeply by extra loads that the amount of energy left is too small to start the car. The car battery monitor described here prevents this problem by disconnecting extra loads in time, ensuring a battery condition which is always good enough for a reliable engine start.

Design by K. Walraven

FORGOTTEN to switch off the car lights? Not to worry, they will go out by themselves, very slowly. Good for your physical shape, too, because the car is bound to need a push-start the next morning.

For campers and caravanners the car battery is a welcome source of energy, because it allows many 12-V operated devices to be used, for instance, a small vacuum cleaner, your mother-in-law's mini hair dryer, a coffee machine, a portable TV, an extra bedside light, an electric shaver, and, of course, the mini stereo rack. Relishing the use of all these luxury devices (which you have become used to at home), no one thinks of recharging the car battery. The joy lasts till the next morning, when the car, after a few feeble and discouraging sounds of the starter engine, refuses to start because of a totally exhausted battery. The only way to get 'on the move' again is to lend some battery power from the neighbours' car, or team up and push-start the car. Obviously, such events are to be avoided at all costs, particularly when you are on holiday.

The preventive action of the car battery monitor described here can be achieved at a very small outlay, and with a small effort as regards construction. The circuit measures the discharge level of the car battery by monitoring the battery voltage. If the voltage is in danger of becoming too low for the car to be started, the extra loads (lights, radio, etc.) are disconnected from the battery. So, if you happen to have forgotten to switch off something, that will not cause a totally drained battery.

The lead-acid battery

Before we tackle the description of the circuit proper, a short description is given of the discharge behaviour of a lead-acid car battery — see **Fig. 1**. Note that the graph shown is typical of a battery that has been in use for some time. The shape of the curve of an old battery is totally different from that produced by a brand new battery.

The graph indicates a relation between the charge condition (ranging from 'discharged' to 'charged') and the open-circuit battery voltage. This feature is exploited by the present battery monitor. If the battery voltage drops below a certain threshold, the load is disconnected to save the remaining energy. The exact value of the threshold differs from battery to battery. Hence, an adjustment is provided in the circuit to enable any threshold to be set, depending on personal requirements.

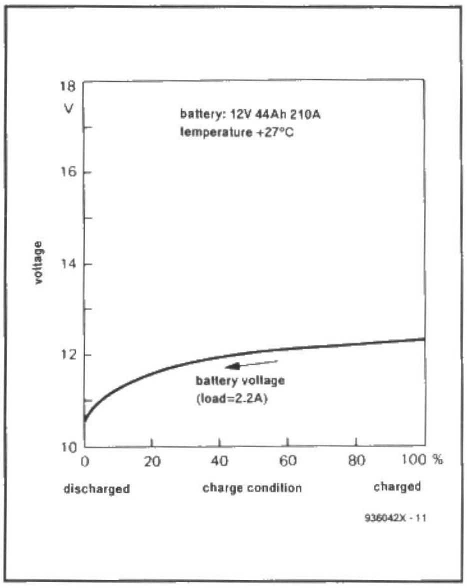
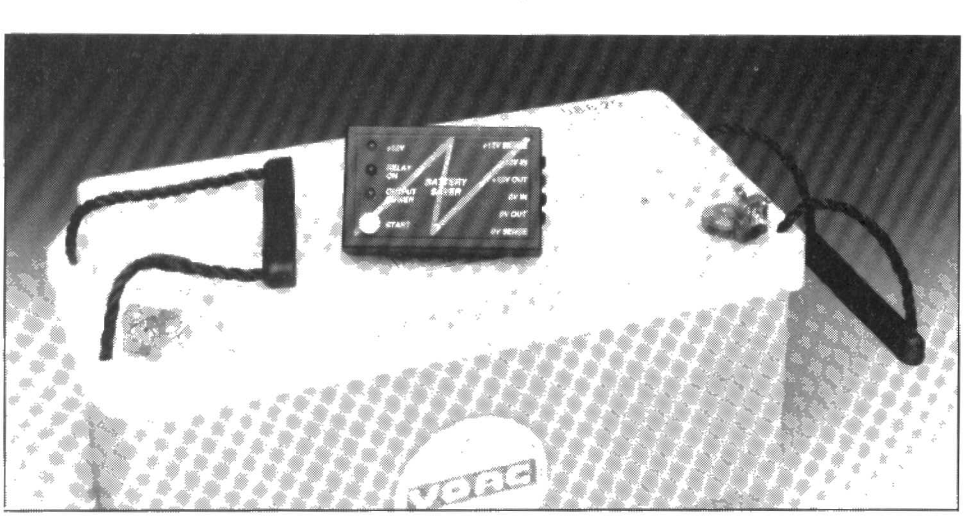


Fig. 1. Typical lead-acid battery discharge graph. The shape of the curve depends on the battery's age and condition.



Circuit description

Looking at the circuit diagram of the battery monitor, **Fig. 2**, you are greeted by an old faithful, the Type 723 voltage regulator in position IC_1. This IC is used to compare a part of the battery voltage with an internal reference voltage. This is done with the aid of an internal opamp. The reference voltage is fixed at 7.15 V, and appears at pin 6. It is applied to the inverting input (pin 4) of the internal opamp. The non-inverting input is connected to the wiper of preset P_1. In this way, a part of the battery voltage appears at pin 5 of the 723. The actual size of the

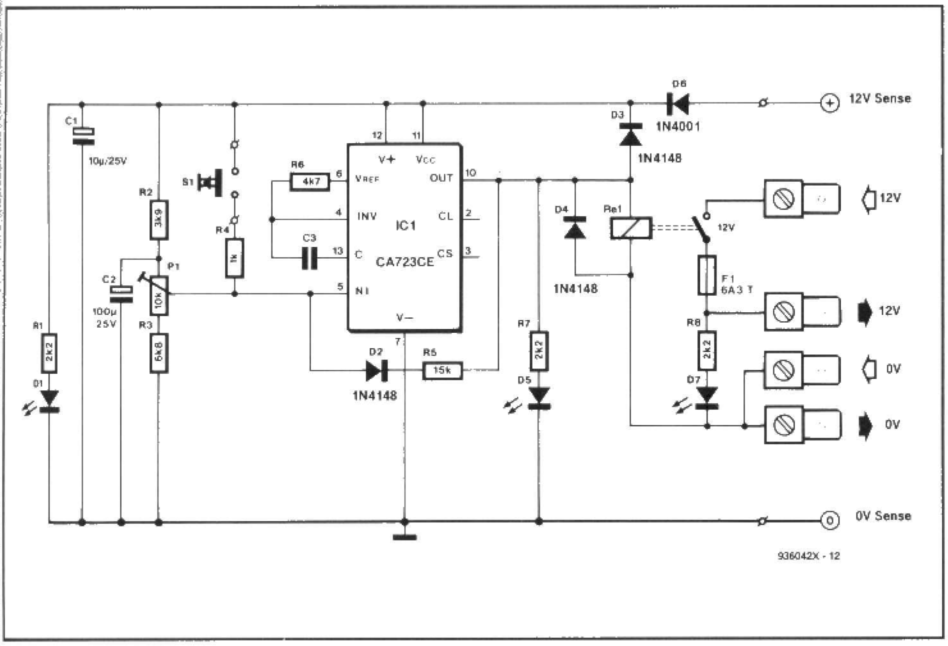


Fig. 2. The battery saver is based on the familiar 723 voltage regulator, which contains, among others, a stable 7.15-V reference and an opamp.

part is determined by voltage divider R_2-P_1-R_3.

The battery voltage measured by the opamp is decoupled by capacitor C_2, which suppresses interference across P_1 and R_3. The output of the internal opamp is connected to pin 10 of IC_1. This output is 'high' if the stepped-down battery voltage at pin 5 is higher than 7.15 V. The opamp output swings 'low' when the voltage at pin 5 drops below the reference level. That also causes the relay, Re_1, to be switched off, and LED D_5 to go out. Diode D_4 suppresses the back-emf voltage generated by the relay coil as the current through it is interrupted. The loads connected to the relay contacts are disconnected from the battery. Our goal has been achieved: if the battery voltage drops below a certain level, a number of loads are switched off automatically via a relay.

Switch S_1 is intended for emergency cases, and allows you to switch the loads on (briefly) after they have been disconnected by the circuit. Fuse F_1 protects the relay contact, the PCB tracks and the connecting wires against too high currents. LED D_7 is the 'fuse intact' indicator. If it does not light while D_5 is on, the fuse is blown.

The use of 'sense' lines requires a separate discussion, which is aided by the drawings in **Fig. 3** and **Fig. 4**. In Fig. 3, the battery monitor 'sees' the voltage across the load. At first glance, this voltage is equal to the battery voltage, but there is a snag. Assuming that the wires between the battery and the load have a resistance of about 0.5 Ω, and the load draws a current of 1 A, the cable 'drops' 0.5 V. The voltage applied to the monitor circuit is, therefore, 0.5 V lower than the battery voltage, causing the load(s) to be disconnected too early. In Fig. 4, the circuit monitors the battery voltage via two sense wires. This setup eliminates the above measurement error, and enables the monitor to keep an eye on the actual battery voltage, ensuring that extra loads are switched off when it is really necessary.

Construction

Although the printed circuit board shown in **Fig. 5** is not available ready-made through our Readers Services, the construction of the battery monitor should not present problems. Alternatively, the circuit may be built on a piece of stripboard. The tracks which carry the load current must be tinned and sufficiently wide. The 'Amp' type spade terminals are secured to the PCB using bolts and nuts (**Fig. 6**). Once the screws have been tightened, the nuts may be soldered to the respective copper areas. The 'Amp' type terminal has three advantages: it is suitable for high currents; it enables an easy connection to clamp-on cable receptacles; and it is widely used for electrical systems in cars.

The IC is fitted in a socket, and the

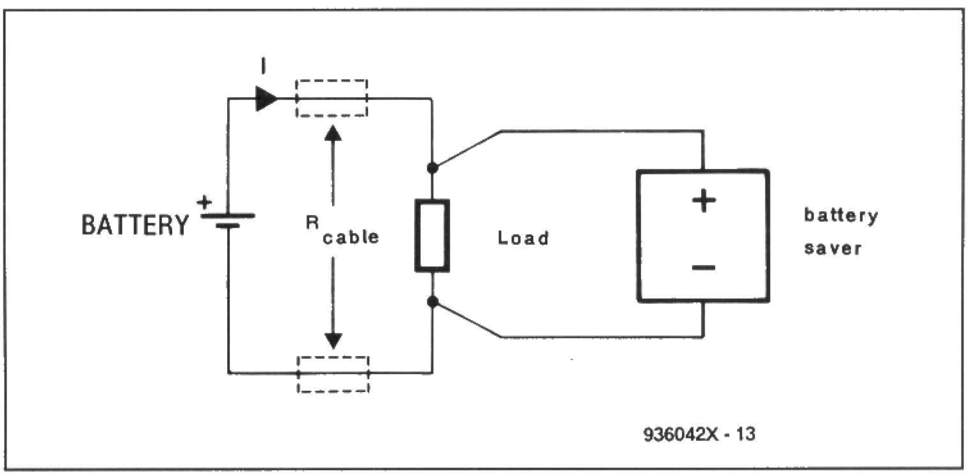


Fig. 3. If the sense wires are connected across the load, the cable resistance may cause the battery monitor to measure a voltage which is lower than that on the battery terminals.

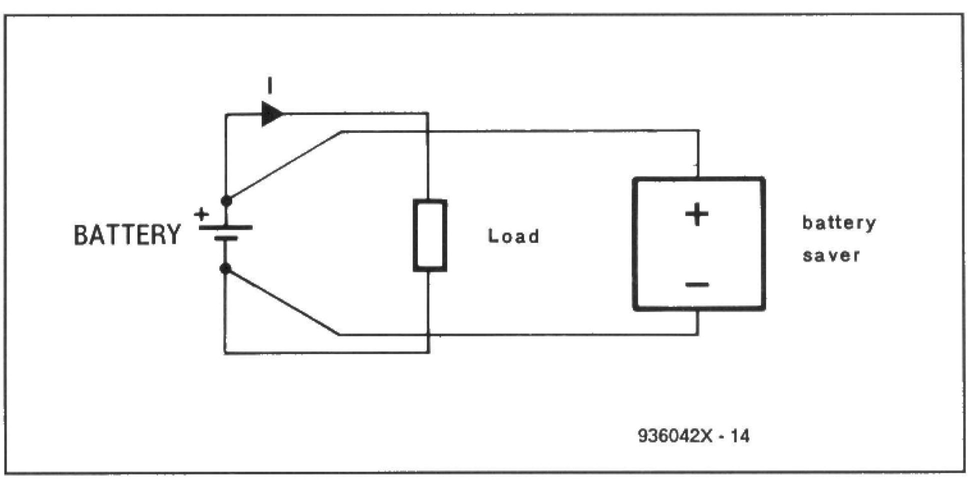
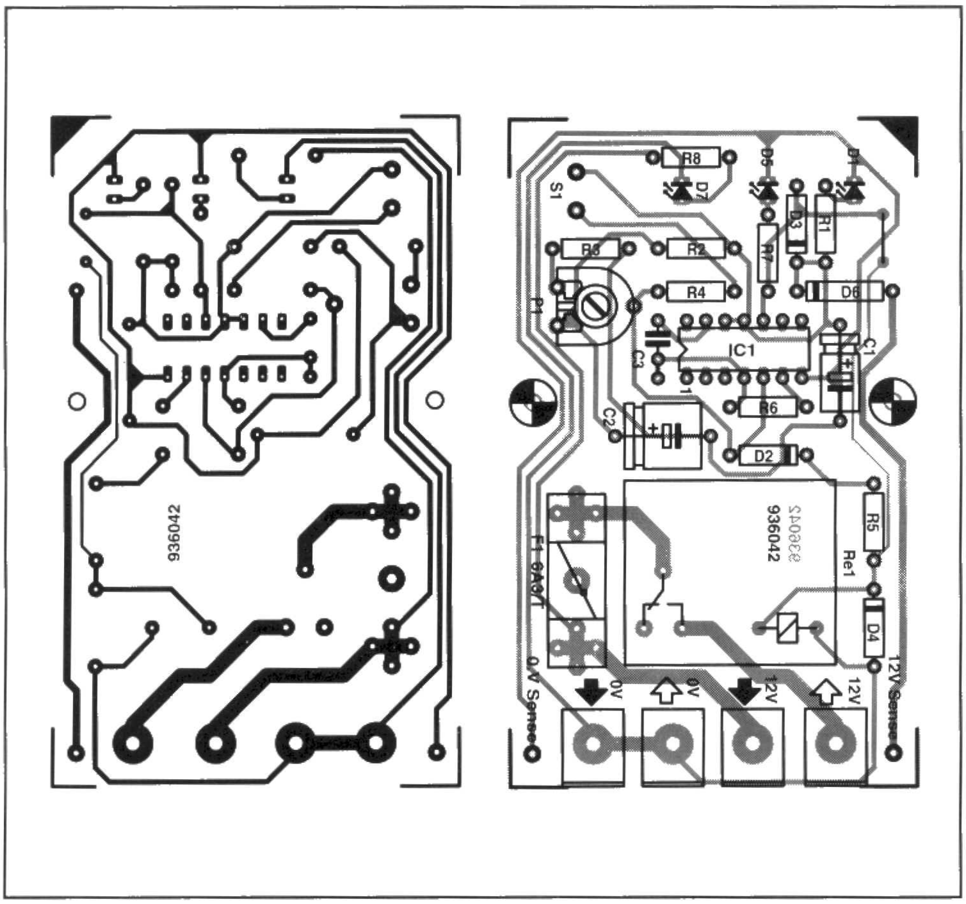


Fig. 4. The proper connection: the sense wires go directly to the battery terminals.



COMPONENTS LIST

Resistors:
R1;R7;R8 = 2kΩ2
R2 = 3kΩ9
R3 = 6kΩ8
R4 = 1kΩ
R5 = 15kΩ
R6 = 4kΩ7
P1 = 10kΩ preset H

Capacitors:
C1 = 10μF 25V
C2 = 100μF 25V
C3 = 1nF
D1;D7 = LED, 5mm, green
D2;D3;D4 = 1N4148
D5 = LED, 5mm, red
D6 = 1N4001
IC1 = CA723CE

Miscellaneous:
S1 = push-to-make button.
Re1 = V23127-A2-A101 (12V; Siemens).
F1 = 6.3A slow
4 'Amp' cable sockets and PCB-mount spade terminals.
Enclosure Pac-Tec HM-kit 6600-902.

Fig. 5. Track layout (direct reading) and component overlay of the PCB designed for the battery monitor (PCB not available ready-made through the Readers Services).

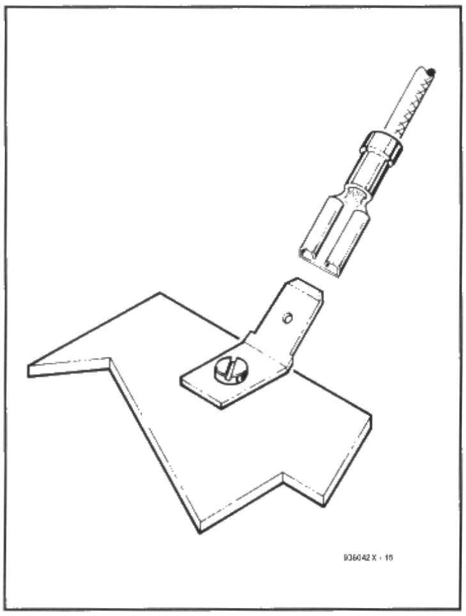


Fig. 6. Illustrating the use of car-type 45° angled 'Amp' terminals and cable receptacles.

mounting of the other parts on the board will not present problems. Do not forget the single wire link on the board, because the circuit does not work without this 'component'.

Once the PCB is fully populated (**Fig. 7**), it may be fitted into a suitable enclosure, using the front panel layout given in **Fig. 8**. Make sure the terminals remain accessible without having to open the case.

To prevent voltage loss in the connecting wires, the circuit is best mounted as close as possible to the car battery. Use heavy-duty copper wire with a cross-sectional area of at least 4 mm^2, or 6 mm^2 if very high currents are switched. Such wire may be obtained from hi-fi shops as loudspeaker cable.

Connect the circuit to the load(s) and the battery as indicated in Fig. 4. If you want the LEDs and the press-key on your car dashboard, they must be connected via wires if the battery

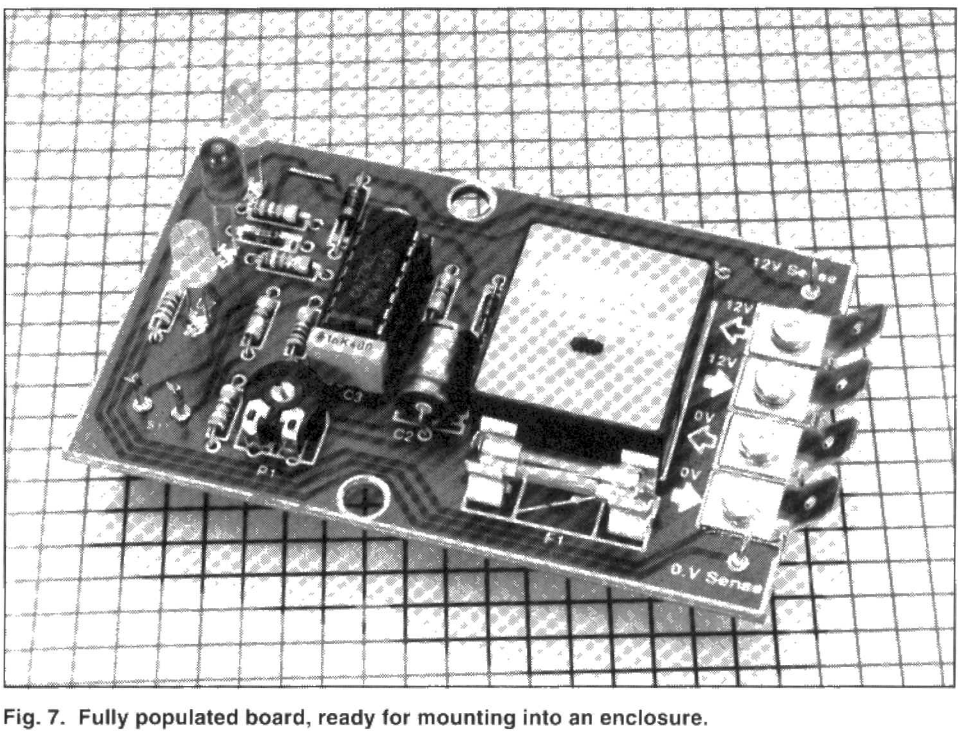


Fig. 7. Fully populated board, ready for mounting into an enclosure.

monitor is mounted close to the battery. Alternatively, if these parts are fitted on the board, the front panel layout shown in **Fig. 8** may be used.

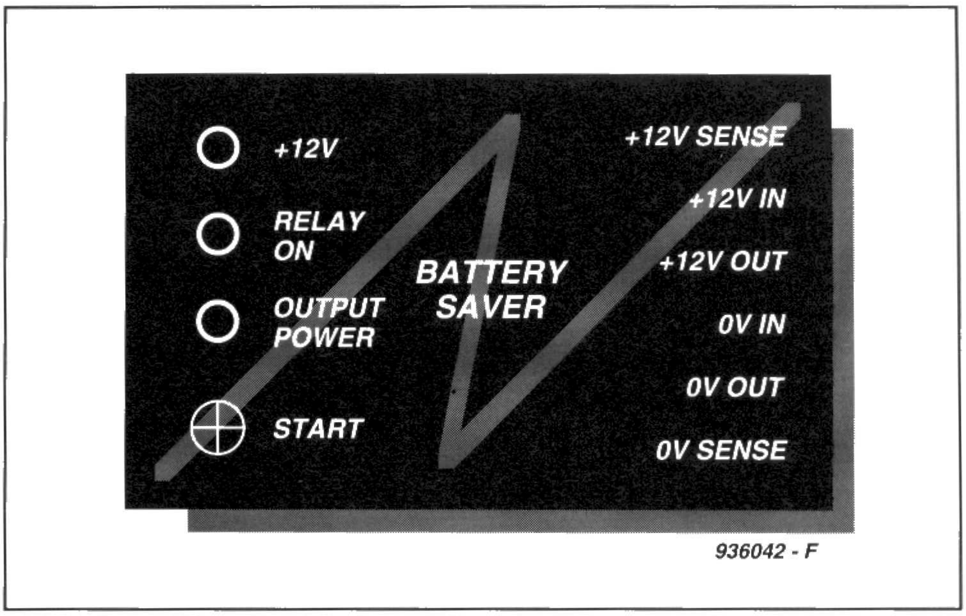


Fig. 8. Suggested front panel layout and lettering (not available ready-made through the Readers Services).

Adjustment

Strictly speaking, preset P_1 can only be adjusted by trial and error. Unfortunately that may take quite a lot of time, whence our suggestion to adjust the circuit as follows.

Connect the circuit to an adjustable power supply, and make the output voltage equal to the desired switching threshold, for instance, 11.5 V or 12 V (guidance values). Adjust P_1 until the relay is just de-energized. If, after some time, it is found that the switch-off voltage is too low or too high, the threshold may be corrected a little by carefully adjusting the preset.

Attention!

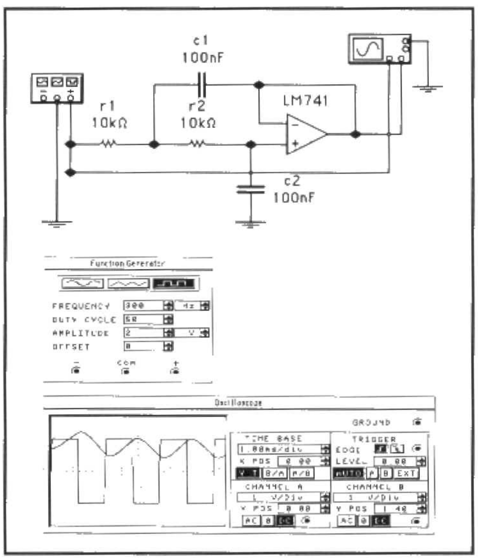
Car batteries are capable of delivering extremely high currents which can cause fire and other hazards. The circuit discussed here is only suitable for 12-V car batteries. For higher battery voltages, for instance, 24 V, the relay must be replaced by a suitable type, and voltage divider R_2-P_1-R_3 requires different resistor values.

(936042)

ELECTRONICS WORKBENCH THE ELECTRONICS LAB IN A COMPUTER

Electronics Workbench (version 3) from Interactive Image Technologies (IIT) is a CAD package specifically geared to rapid design and verification of analogue and digital circuits. Since the circuit is simulated on a computer, making changes to it is much quicker than the traditional breadboard method, i.e., scouring for components, fitting them, soldering and connecting the test equipment after every change. None the less, no electronics design program, including EWB, does away with the need to build a prototype from real components, since the proof of the pudding is in the eating. EWB should be of great interest to students, as well as to the reasonably advanced hobbyist. Once a circuit is known to function properly using EWB, you have the green light to actually build it.

Drawing a circuit in EWB is helped by a 'parts bin' containing most commonly used electronic parts. Although the drawing phase is pretty fast and efficient, producing the circuit response (measured by various instruments, including an oscilloscope and a bode plotter) is sluggish. Tests were carried out on a Macintosh IIci computer. Digital circuits, by contrast, produced test results much faster.



Adding parts to the library is easy if you duplicate the existing 'ideal' version of a particular component, and then assign parameter values to the 'real' part to be added. From then on, the part is available for all new work.

Although the components pick and place operations work fine, the program sometimes has a quirk in the way in which the parts are actually positioned and connected. Also, although they are electrically correctly connected, the leads to the test bench instruments sometimes take surprising paths criss-cross through the circuit.

Among the shortcomings of the program that should be mentioned are the inability to read netlists and so put circuits generated by other programs on the test bench, and the unusually simple shapes assigned to some ICs like the 555. Also, it would be desirable to have a rather more elaborate parts library, and to be able to mix linear with digital electronics. Fortunately, a library extension (Model Set 1), and a set of 150 'standard circuits' can be obtained from IIT.

None of these criticisms, however, can alter my opinion that Electronics Workbench is well worth considering if you are after a drastical reduction of cost and time spent on the design of analogue and digital circuits of up to medium complexity. *tech. ed.*

Electronics Workbench costs £199 (excl. VAT and p&p), and is available for DOS, Windows and Apple platforms, from **Robinson Marshall (Europe) Ltd., Nadella Building, Progress Close, Leofric Business Park, Coventry CV3 2TF. Telephone: (0203) 233216, fax: (0203) 233210.**

GENERAL-PURPOSE INFRA-RED VOLUME CONTROL

Home made audio amplifiers are rarely equipped with a remotely controlled motor-driven volume control. Curious, because motor-driven pots are high-quality products, and consequently should appeal to the 'high-end' audio fraternity. Fortunately, prices of motor-driven volume controls have come down to acceptable levels, while infra-red remote controls can be obtained fairly easily from TV spare parts sources. Time to put two and two together.

Design by T. Giesberts

THE number of electronics hobbyists interested in building audio amplifiers and related items is traditionally fairly high. It may therefore safely be assumed that a fair number of readers have been waiting for a project of the kind described here.

A relatively simple circuit is discussed that enables the volume of, for instance, a home-made amplifier or a similar circuit (Audio DAC, Ref. 1) to be controlled via an ordinary infra-red remote control. The remote control unit may be any type, as long as it is RC5 compatible. The self-learning (or 'intelligent') brand offered by some radio and TV retailers may also be used.

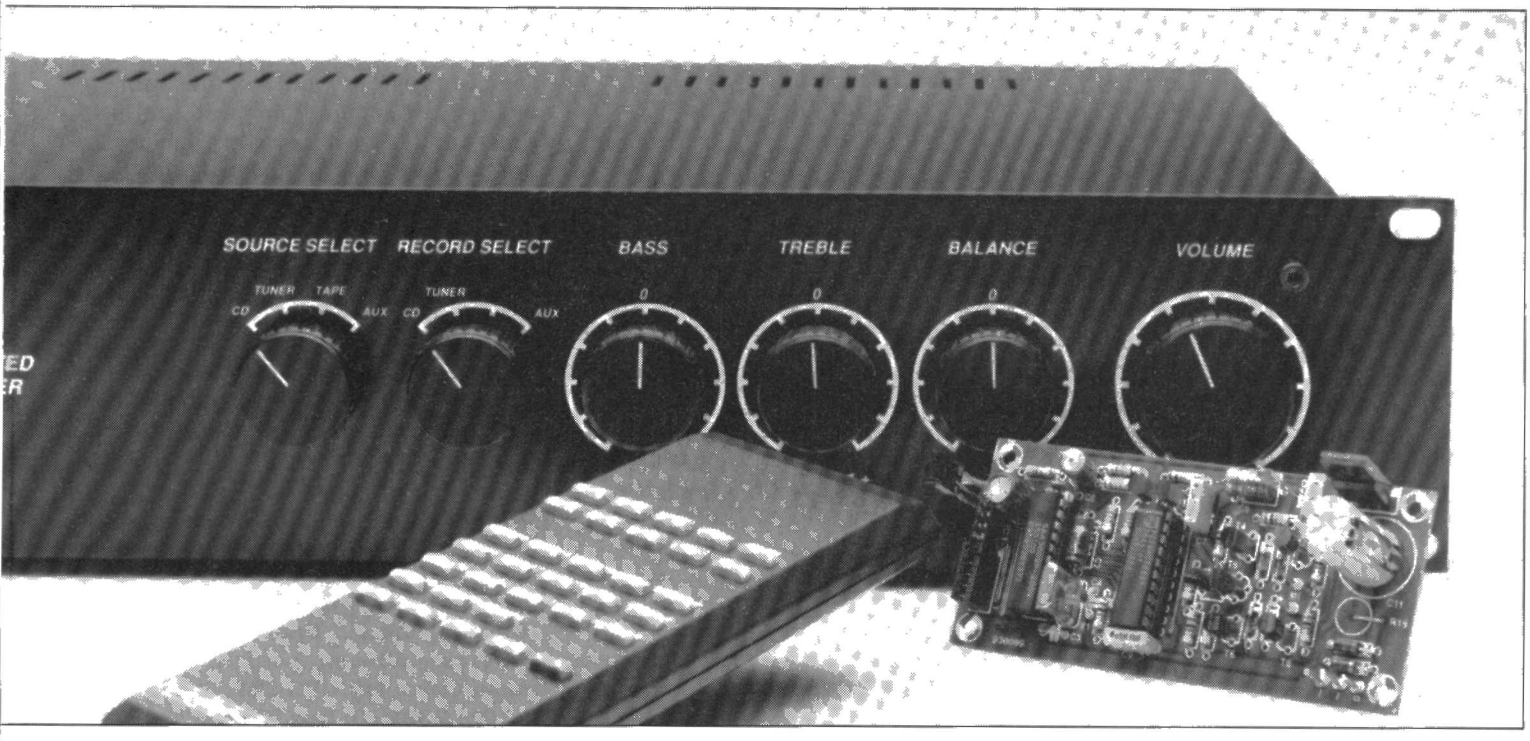
Thanks to the use of integrated circuits, the number of components in the project is small, while motor-driven potentiometers are available at reasonable prices these days. All in all, many high-end audio enthusiasts will find the remote-controlled volume control a worthwhile upgrade for an existing preamplifier.

The drive circuit

The circuit diagram of the infra-red volume control is shown in **Fig. 1**. The circuit consists mainly of three ICs for receiving and decoding the infra-red signal, and a couple of transistors to drive the motor. The circuit has its own power supply, which is fed by the secondary winding of the mains transformer used in the preamplifier. That is done to save you the cost of an extra transformer, while the preamplifier supply will happily supply the small extra current drawn by the volume control.

The infra-red signal transmitted by the remote control unit is picked up by IC_1. The IS1U60 from Sharp not only contains an infra-red sensitive diode, but also a complete receiver consisting of an amplifier, a limiter, a 38-kHz pass-band filter, a demodulator and an output stage — all in a case with a size of only a few millimetres. Furthermore, the IS1U60 is fully screened against electromagnetic interference, and features a lens element to bundle and focus infra-red light. Sharp Electronics in their datasheets specify a minimum operating range of 5 metres, while 3 metres is guaranteed at a horizontal angle of incidence of 30°, or a vertical angle of incidence of 15°. In practice, the range achieved by the receiver is much larger — our prototype easily covered 15 m. The high sensitivity of the IS1U60 makes the device just the thing for the present application.

The received and filtered drive signal may be applied directly to the decoder, a SAA3009 (IC_2). The SAA3009 is related to the SAA3049, which was used in earlier projects in this magazine, including a multi-purpose RC5 infra-red receiver (Ref. 2). The main difference between the SAA3009 and the SAA3049 is the higher output current capacity of the former: 10 mA



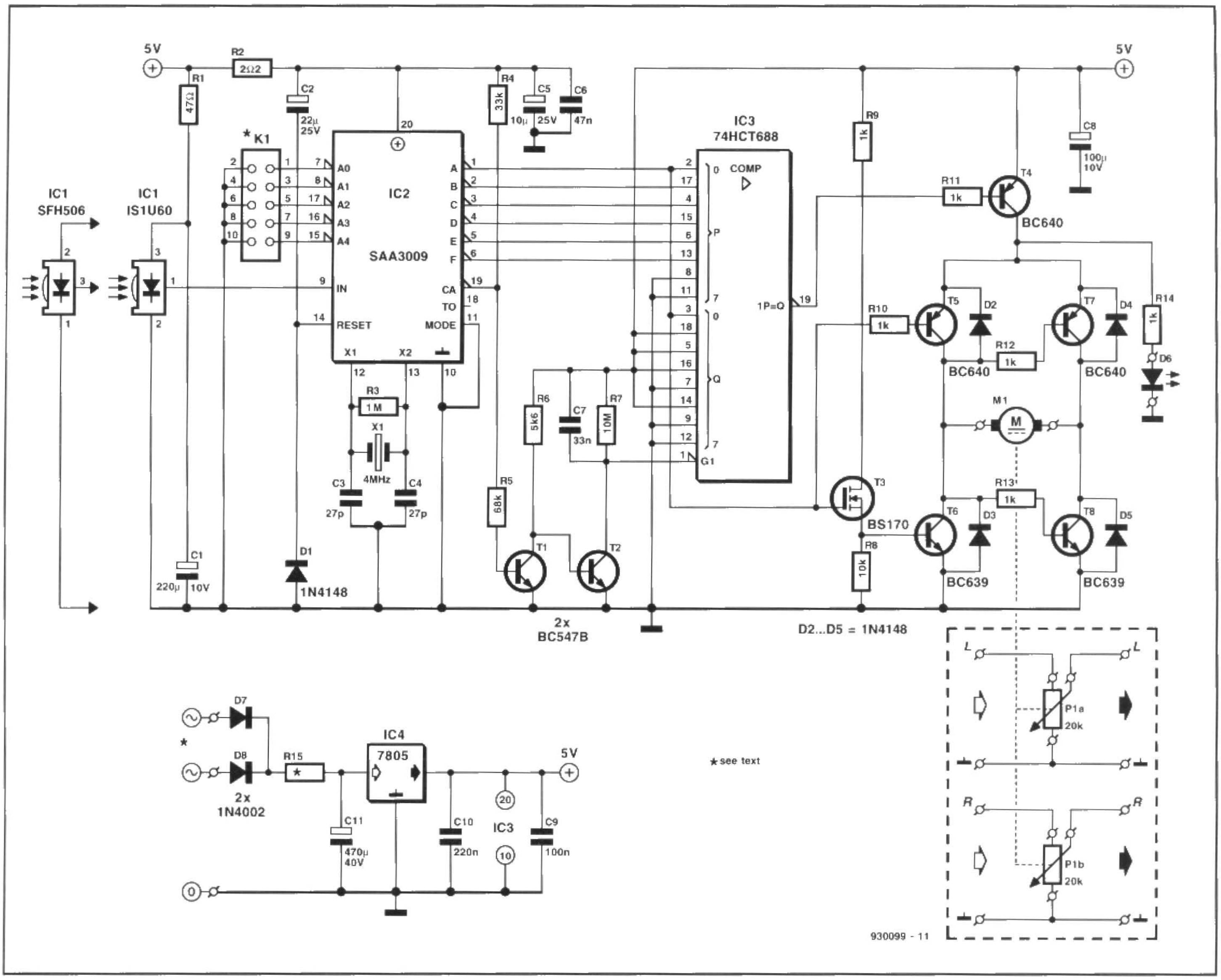


Fig. 1. The circuit of the IR volume control consists of three sections: an infra-red receiver, a decoder and a driver for the motorized potentiometer.

against 3 mA for the SAA3049. Obviously this is reflected by the 3009's total current consumption, which can go up to 70 mA.

The function of the SAA3009 is to convert the received data into a binary code. The serial RC5 signal contains two important data: the system address and the command proper. In the RC5 set of codes, the function 'preamplifier' is assigned system address 16. As a matter of course, every user is free to assign a different address to the amplifier (or, in more general terms, the volume control), if '16' is already in use, or if the potentiometer is fitted in a circuit which is not an amplifier. The address is set with the aid of inputs A0-A4 on the SAA3009. Address '16' is selected by making A0 through A3 logic low (jumpers fitted), while no jumper is fitted for A4. The address reserved for TV sets is '0', which is selected by fitting all four jumpers on jumper block K_1.

The SAA3009 operates in single-system mode (SSM), which means that it responds to only one address. Short-circuiting resistor R_5 causes the IC to switch over to combined system mode (CSM). The decoded address is then available on pins A0 through A4, which then function as 'active-low' outputs.

In the RC5 protocol, the system address is followed by the command proper. That is decoded in IC_2, and made available at outputs A through F. In principle, the IC is capable of decoding up to 64 commands in addition to the 32 system addresses. In the present circuit, however, only two commands matter:

address	F	E	D	C	B	A	Command
16	0	1	0	0	0	0	vol. up
17	0	1	0	0	0	1	vol. down

Outputs A-F are connected to the 'P' inputs of a digital comparator, IC_3. The output of this comparator, labelled P=Q, goes low when the dataword at the 'P' inputs matches the dataword at the 'Q' inputs. That happens when P=0010111. The least significant bit, A, forms the only difference between the 'volume up' and 'volume down' commands. This bit is connected to the P0 and the Q0 pin of the comparator, to make sure that they are always at the same level. It also drives the gate of MOSFET T_3 and so determines the left/right rotation of motor M_1.

If the dataword at the P inputs matches the preset dataword at the Q inputs, T_4 is driven, and the motor driver is powered. Next, the level of the 'volume up/down' bit, A, determines the rotation direction of the motor. If A=0, T_5 conducts. Consequently, T_3 is switched off, so that the base of T_6 is pulled to ground, causing this transistor to be switched off also. However, since a voltage of 5 V is present on the collector of T_6, T_8 is switched on, and T_7 is switched off. The upshot is that a current flows through the motor winding, via T_4, T_5, T_8, and into the ground rail.

If A=1, the above state is reversed: T_5 is then off, so that T_3 and T_6 are switched on. T_6 pulls the base of T_7 to ground, so that T_7 starts to conduct also. Consequently, current flows in the opposite direction through the motor winding, via T_4, T_7, and into the ground rail again, this time via T_6. In this way, the level of bit A determines the direction of the motor. LED D_6 lights any time the motor operates.

The sub-circuit around T_1 and T_2, between the CA output of IC_2 ('command received', pin 19) and the enable input (G) of IC_2, affords additional protection against noise, and also serves to detect when a new system address is received. If a system address is decoded, the CA output supplies a rectangular signal which is low for 15 ms, and high for 105 ms. Because of the large time constant of network R_7-C_7, transistor T_2 operates as a rectifier on this signal. Its collector will therefore remain logic high as long as the signal is present at pin 19. As soon as the signal disappears, IC_3 is immediately disabled.

Immediately after the supply is switched on, IC_1 is reset via capacitor C_2. Diode D_1 ensures that C_2 is rapidly discharged when the supply is switched off. The toggle bit (pin 18), which changes state on every command received, is not used in this circuit.

The power supply is conventional. Diodes D_7 and D_8 rectify the alternating voltage supplied by the secondary winding of the preamplifier's mains transformer, and C_{11} smooths the direct voltage. Resistor R_{15} reduces the regulator input voltage so that the dissipation of the 7805 is kept within specifications. The value of the 5-watt resistor depends on the transformer's secondary voltage. For example, if the secondary voltage is 30 V, the rectified voltage will be of the order of 40 V. Since the maximum input voltage of a 7805 regulator is 35 V, resistor R_{15} will have to drop at least 5 V. The current consumption of the IR volume control being about 50 mA, R_{15} then takes a value of 5 V/50 mA = 100 Ω.

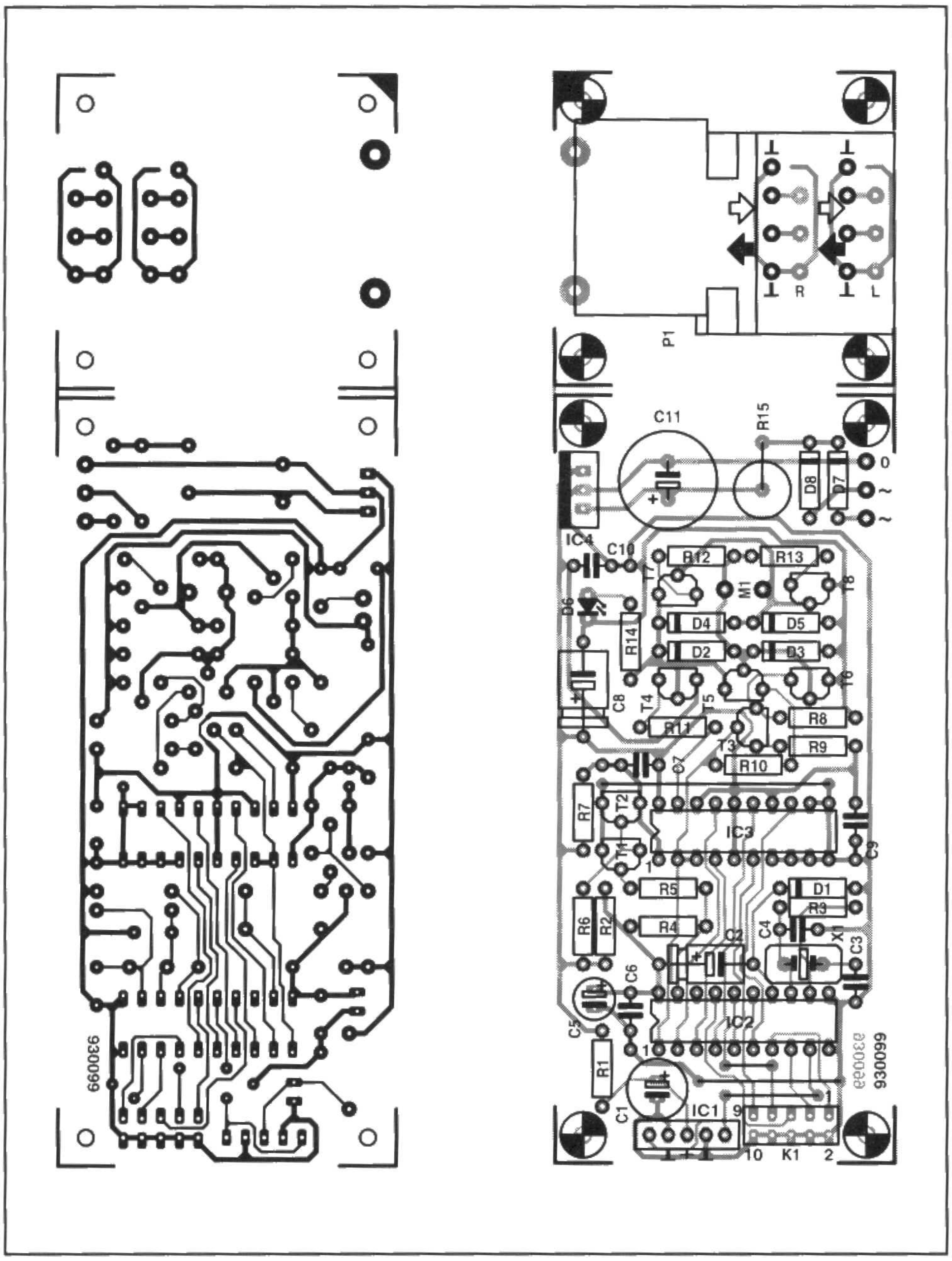


Fig. 2. The PCB may be cut into two, separating the driver board from the potentiometer board. The IR receiver circuit, IC1, and the potentiometer are best mounted onto the preamplifier's front panel. This printed circuit board is available ready-made through the Readers Services (see page 110).

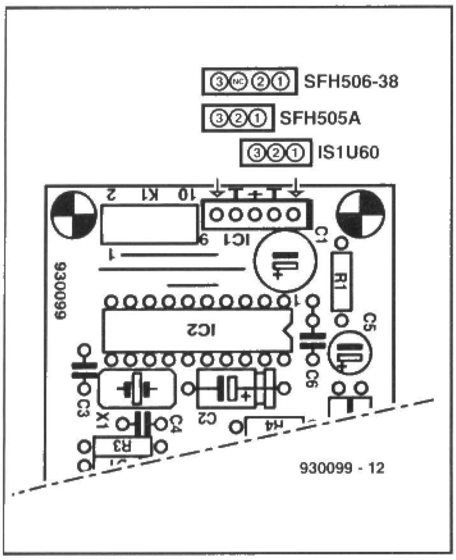
COMPONENTS LIST

Resistors:
R1 = 47Ω
R2 = 2Ω2
R3 = 1MΩ
R4 = 33kΩ
R5 = 68kΩ
R6 = 5kΩ6
R7 = 10MΩ
R8 = 10kΩ
R9-R14 = 1kΩ
R15 = see text
P1 = 20kΩ log. stereo potentiometer w. motor drive (Alps type 20KAX2)

Capacitors:
C1 = 220µF 10V radial
C2 = 22µF 25V
C3;C4 = 27pF
C5 = 10µF 25V radial
C6 = 47nF ceramic
C7 = 33nF
C8 = 100µF 10V
C9 = 100nF
C10 = 220nF
C11 = 470µF 40V

Semiconductors:
D1-D5 = 1N4148
D6 = low-current LED
D7;D8 = 1N4002
T1;T2 = BC547B
T3 = BS170
T4;T5;T7 = BC640
T6;T8 = BC639
IC1 = IS1U60 (Sharp), or SFH506-38 (Siemens)
IC2 = SAA3009
IC3 = 74HCT688
IC4 = 7805

Miscellaneous:
K1 = 10-way boxheader w. 5 jumpers.
X1 = 4MHz quartz crystal
Printed circuit board 930099 (see page 110).



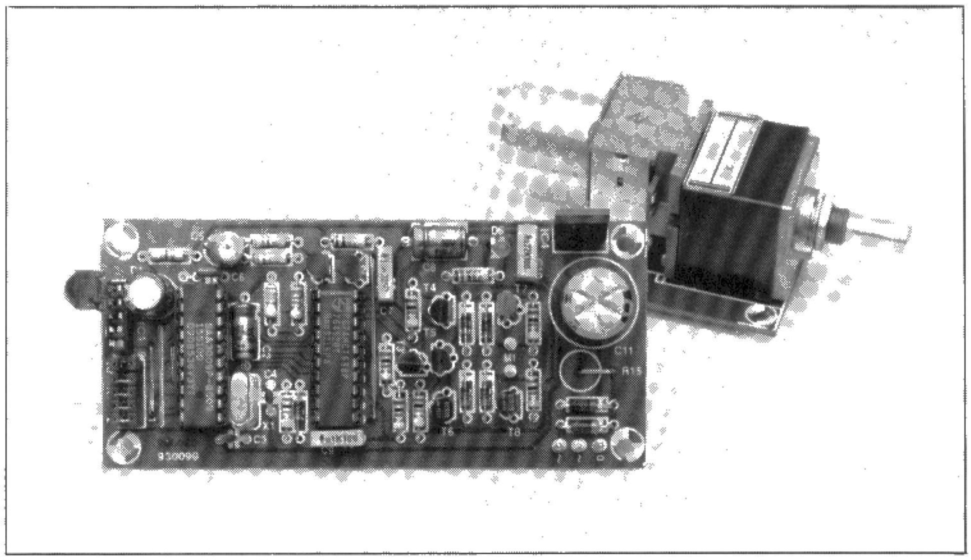


Fig. 3. **Three different types of infra-red receiver IC may be used with the wireless volume control. Note the different pinnings, which are catered for by a 5-way SIL socket on the driver board.**

Construction

The printed circuit board of which the artwork is shown in **Fig. 2** consist of two sections which are separated with a jigsaw. The combined board is available ready-made through the Readers Services. The larger board contains the drive electronics, while the smaller one serves to mount the motor-driven potentiometer. Solder pins may be fitted at the solder side of the potentiometer board. The small board is useful if you want to fit the pot at the back of the preamplifier case. The PCB is not strictly necessary if you decide to secure the pot behind the front panel, in which case the wires may be soldered directly to the tags on the motor.

As long as you do not forget to fit the wire links, and take care to fit the polarized components the right way around, no special problems should be encountered in the construction. Work carefully, and follow the parts list and the component overlay.

An alternative for the IS1U60 is the Siemens SFH506-38. As shown in **Fig. 3**, the two ICs have different pinnings, so take care when connecting one to the input of the control board. The SFH505A appears to be obsolete, and should not be used in conjunction with the SAA3049.

Once populated, the circuit board may be fitted into the preamplifier enclosure. Replacing the existing potentiometer in the preamplifier with the motorized type is fairly easy in most cases. The PCB with the drive electronics is best fitted behind the preamplifier's front panel, so that the infra-red diode contained in IC_1 can 'look out' through a hole. If space is tight, the PCB may have to be located a little further away from the front panel. In that case, IC_1 may be connected to the board via a short length of screened cable.

Final notes

Those of you who are wondering how to program an existing remote control unit for system address 16 may find the information in Ref. 2 of use, since modification details are given in that article.

Finally, just before the present design was finalized for publication, information was received from Philips Components that the SAA3009 is supplied in limited quantities only. A suitable equivalent is the SAA3049, which does not suffer from supply problems. However, the use if the 3049 does require a number of resistors to be added, as shown in **Fig. 4**. Eleven 10-kΩ pull-up resistors are fitted at terminals A0-A4 and A-F of IC_2, and a 68-kΩ resistor in parallel with D_1. Components R_3, C_3, C_4 and the quartz crystal may be omitted.

(930099)

References:

1. AF digital-to-analogue converter, parts 1, 2 and 3, *Elektor Electronics* July/August, September and October 1992.

2. Universal RC5 code infra-red receiver, *Elektor Electronics* January 1992.

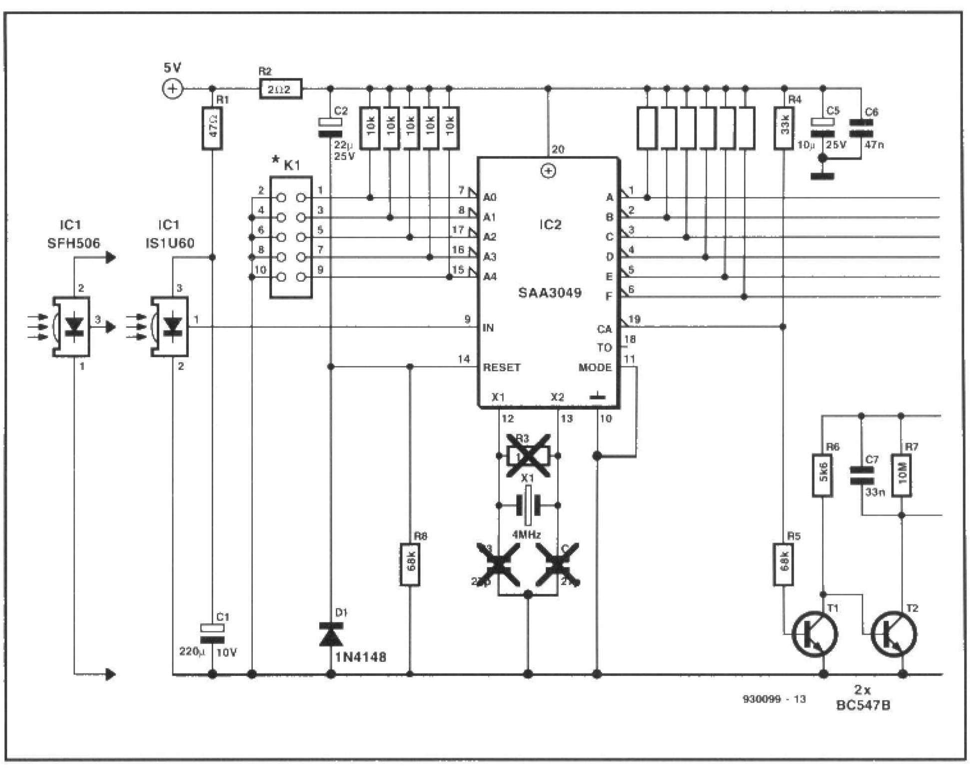


Fig. 4. **If a SAA3049 is used instead of a SAA3009 you have to add 12 resistors.**

FIGURING IT OUT

PART 18 – THE LAPLACE TRANSFORM

By Owen Bishop

This series is intended to help you with the quantitative aspects of electronic design: predicting currents, voltage, waveforms, and other aspects of the behaviour of circuits. Our aim is to provide more than just a collection of rule-of-thumb formulas. We will explain the underlying electronic theory and, whenever appropriate, render some insights into the mathematics involved.

Systems of many kinds, from electronic circuits to the motion of planets and the spending patterns of supermarket customers, may be modelled with differential equations. Setting up the equations is usually fairly easy. Solving the equations may be unexpectedly dfficult. Obviously, a solution may be more readily reached if we assume that certain values are zero or remain constant, but this may make the model unrealistic, or at least limit its applicability. Various techniques have been devised to assist the solution of differential equations. One of these is the operator **D**, described in Part 14. Another technique, known as Euler's method, is easy to use but, since it relies heavily on approximations, is not entirely reliable. A third and powerful technique is the Laplace transform.

Transforms

Given a set of values, we transform them by performing a specified mathematical operation on each of them. A well-known example is the logarithmic transform. Take this set of numbers:

0.5, 2, 3.5, 7, 20.22.

The logarithmic transform of these is obtained by taking the natural logs, or logs to any other base, if preferred. Natural logs can be found with a pocket calculator and we obtain the transformed set (shown here to 4 decimal points):

–0.6931, 0.6931, 1.2528, 1.9459, 3.0067.

The members of the original set map in a one-to-one manner with the members of the transformed set. We can also perform the inverse transform, using the e^x function on the calculator, and recover the original set.

At one time, logs were used extensively to assist in calculations, either in the form of log tables, or on the slide rule. This is because they have the property that adding logs is the equivalent of multiplying ordinary numbers. For example:

$$\ln 2 + \ln 3.5 = 0.6931 + 1.2528 = 1.9459 = \ln 7.$$

Taking logs, adding them, and taking the antilog of the sum, has given us the product. Adding is easier than multiplying, especially when there are many decimal places.

Finding a fractional power of a number is a troublesome operation. For example, find the value of $3.5^{2.4}$. The log transformation comes to our aid because multiplying a log by a number is the equivalent of finding the power. In this example:

$$2.4 \times \ln 3.5 = 2.4 \times 1.2528 = \ln 20.22$$

that is,

$$3.5^{2.4} = 20.22.$$

Multiplying the log, then finding the antilog, has given us the power. Multiplying is easier than taking powers, especially when they are fractional. Summing up, we use the log transform because it makes certain maths operations easier to do. We use the Laplace transform for the same reason.

Laplace transform

The Laplace transform operates on a time function. By this we mean that a quantity (such as voltage) is specified by a function $f(t)$, in which time is the independent variable. For example, $u = 3 \sin \omega t$. We say that u is in the **time domain**. We sometimes write $u(t)$ and $i(t)$ for the function instead of $f(t)$, the lower case letter indicating the quantity involved. The Laplace transform of a time function $f(t)$ is $F(s)$, where:

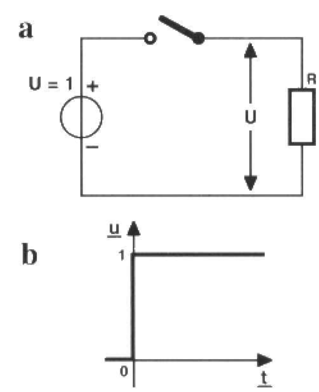
$$F(s) = \int_0^\infty f(t)\,e^{-st}\,dt \qquad \text{[Eq. 132]}$$

The main condition attached to this transform is that the integral must be **convergent**, which means that the integral must approach a definite limiting value as t becomes large, and not become infinitely large itself. Most of the time, functions met in electronics conform to this requirement. It is also specified that $f(t) = 0$ when $t = 0$. If it is not, we insert starting conditions into the equations.

Unit step function

The Laplace transform is not unduly complicated when the time function is a simple one. As an example of the transform, we look at the unit step function, which, in effect, is equivalent to just turning on the power switch—see **Fig. 147a**. Up to the instant of the starting time, the switch is open. The voltage across the resistor is $u = 0$. The switch is closed when $t = 0$ and the voltage across the resistor instantly rises to $u = 1$. We have a unit step—see **Fig. 147b**. The function describing this is piecewise:

$$u(t) = 0 \qquad -\infty < t < 0^-$$



930010-XVIII-147

Fig. 147

$$u(t) = 1 \qquad 0^+ < t < +\infty.$$

Note that we specify u up to the instant **before** the switch is closed ($t = 0^-$) and from the instant **after** the switch is closed ($t = 0^+$). It is undefined for the instant of closing. The second of the two functions above is the one we transform. Substituting in **Eq. 127**:

$$U(s) = \int e^{-st}\,dt$$

$$= \frac{-1}{s}[e^{-st}]_0^\infty = 1/s$$

As t approaches infinity, e^{-st} approaches zero; when t is zero, $e^{-st} = e^0 = 1$. The integral reduces to $1/s$. Similarly, if the step is other than a unit step, for example u = a, the transform is a/s.

Other transforms

In a similar way, we can find the transforms of other common functions. As might be expected, the integrations required are more complicated than that for the simple step function above. However, there is no need to worry about this as the transforms of

all the most frequently met functions are available in a table (see Box 1). The transforms in Box 1 are mostly needed when we are performing the reverse transform, as we shall see later. Since our calculations usually begin with a differential equation, we need to know how to transform these. Box 2 shows the transforms of the terms of a differential equation and the way to use these and the transforms in Box 1 is explained in the example below.

A differential equation consists of several terms, usually including a constant, a first differential (dy/dt), a second differential (d^2y/dt^2), and possibly differential terms of third or higher orders. An example taken from Part 13 is a model of a series *LCR* circuit in which $R = 500\ \Omega$, $C = 2\ \mu F$, and $L = 100$ mH. The differential equation, based on **Eq. 94**, is:

$$d^2i(t)/dt^2 + 5000\,(di(t)/dt) + (5\times10^6)i(t) = 0 \qquad \text{[Eq. 133]}$$

Note that we have replaced i in **Eq. 94** with $i(t)$ to emphasise that i varies in the time domain – it is a function of t – but it represents exactly the same quantity. Our aim is to discover the way i varies with time. In other words, to obtain the function $i(t)$. To solve this equation, we will first find its transform. It can be shown that, if we find the transform of each term individually and sum the transforms for each side of the equation, this gives the transform of the whole equation. Working from right to left:

- The transform of 0 is 0 (a unit step in which a = 0 – see Box 1)
- The transform of $(5\times10^6)i(t)$ presents a problem because, until we have solved the equation, we do not know the form of the function i. Represent it by a symbol, using the corresponding capital letter, and say that its transform is $I(s)$. Note that the transform is in the s domain, which can be shown to be the **frequency domain**. Since any constant multiplier also multiplies the transform, the transform of $(5\times10^6)i(t)$ is $(5\times10^6)I(s)$.
- The transform of $5000(di(t)/dt)$ can be obtained by integration according to **Eq. 132**. Skipping over the intermediate stages, which are a conventional integration by parts, we find that the transform of $di(t)/dt$ is $sI(s)-i(0^+)$.

As above, we have had to state the transform in terms of $I(s)$, because we do not yet know the form of $i(t)$. The term $i(0^+)$ is the current flowing at the instant that timing begins. Again, we multiply the transform by the coefficient of the original term, so the transform of $5000[di(t)/dt]$ is

$$5000sI(s) - 5000i(0^+).$$

In Part 13, we stipulated that the current is 2 mA when timing begins, so $i(0^+) = 0.002$ and the transform is $5000sI(s)-10$. Note that instead of having to differentiate $i(t)$, we have simply multiplied its transform by s to give $sI(s)$. Multiplying by s is much easier than differentiating, which is the reason for using the transform.

- The transform of the extreme left-hand term, $d^2i(t)/dt^2$, is also made according to **Eq. 132** and, assuming that no current is flowing at the instant the power is switched on, it produces

$$s^2I(s)-si(0^+)-di(0^+)/dt.$$

The second differential is obtained by multiplying $sI(s)$ by s; as before, it is much easier to multiply by s than it is to differentiate. In this expression, $i(0^+) = 0.002$, as above. $di(0^+)/dt$ is the rate of change of current when timing begins. In Part 13, we said that this is 0.05 A s^{-1}, so the tranform under these starting conditions becomes

$$s^2I(s)-0.002s - 0.05.$$

Summing the transforms of the terms on each side of **Eq. 133**, we obtain a new equation in the frequency domain:

$$s^2I(s)-0.002s-0.05 + 5000sI(s)-10 + (5\times10^6)I(s)=0$$

The next step is to simplify this in order to find $I(s)$, the Laplace transform of **Eq. 133**. Examination shows that $I(s)$ is a factor in three of the terms. The equation becomes:

$$\{s^2+5000s+(5\times10^6)\}I(s) = 0.002s+10.05$$

Function	**Transform**	**Conditions**
1 (unit step function)	$1/s$	$\Re(s) > 0$
a (step function)	a/s	$\Re(s) > 0$
t (ramp function)	$1/s^2$	
e^{at} (growth function)	$1/(s-a)$	$\Re(s) > a$
e^{-at} (decay function)	$1/(s+a)$	$\Re(s) > -a$
$1-e^{at}$	$a/s(s-a)$	$\Re(s) > a$
$1-e^{-at}$	$a/s(s+a)$	$\Re(s) > -a$
$\sin \omega t$	$\omega/(s^2+\omega^2)$	$s > 0$
$\cos \omega t$	$s/(s^2+\omega^2)$	$s > 0$

$\Re(s)$ is the real part of s in those cases where s is a complex number.
a and ω are constants.

Box 1 – Laplace transforms of functions.

Term	**Transformed term**
Constant a	a/s
Function $f(t)$	$F(s)$
First derivative $f'(t)$	$sF(s)-f(0^+)$
Second derivative $f''(t)$	$s^2F(s)-sf(0^+)-f'(0^+)$
Integral $\int_0^t f(t)\,dt$	$F(s)/s+f(0^+)/s$

Box 2 – Laplace transforms of differential equations.

1. Factorise the denominator of the original fraction, if possible.
2. Match the factors against one or more of the formats shown in Box 4.
3. Box 4 sets out the form of the partial fractions; write them out as an identity.
4. Clear fractions by multiplying both sides of the identity by the original denominator.
5. Equate coefficients of each power of x, obtaining equations for constants A, B, etc.
6. Solve these equations to find A, B, etc.

Example: Express $(3x+5)/(x^2-x-12)$ as a partial fraction.

Step 1: $(3x+5)/(x^2-x-12) = (3x+5)/(x+3)(x-4)$.

Step 2: There are two factors, both have the form $(x+a)$. So, there are two partial fractions, both with the form $A/(x+a)$.

Step 3: $(3x+5)/(x+3)(x-4) \equiv A/(x+3)+B/(x-4)$.

Step 4: $3x+5 \equiv A(x-4)+B(x+3)$
$\equiv Ax-4A+Bx+3B$.

Step 5: Equating coefficients of x: $A+B = 3$.
Equating constants: $-4A+3B = 5$.

Step 6: $A = 4/7$ and $B = 17/7$; partial fractions are:
$(3x+5)/(x^2-x-12) = 4/7\,(x+3)+17/7\,(x-4)$.

Further worked examples appear in the text.

Box 3 – Partial fractions.

Factors in denominator	**Partial fractions**
$x+a$	$A/(x+a)$
x^2+ax+b	$(Ax+B)/(x^2+ax+b)$
$(x+a)^2$	$A/(x+a)$ and $B/(x+a)^2$
$(x+a)^3$	$A/(x+a)$, $B/(x+a)^2$ and $C/(x+a)^3$
$(x^2+ax+b)^2$	$(Ax+B)/(x^2+ax+b)$ and $(Cx+D)/(x^2+ax+b)^2$

Box 4 – Formats for partial fractions.

so that:

$$I(s)=\frac{0.002s+10.05}{s^2+5000s+(5\times10^6)}$$

The expression below the line can be factorised, using the quadratic equation formula (available on many scientific calculators) to give:

$$I(s)=\frac{0.002s+10.05}{(s+1382)(s+3618)}$$

[Eq. 134]

We are working toward getting the terms of the equation into the same form as one or more of the transformed expressions in Box 1. Many of these have a single factor, such as $(s–a)$, beneath the line. The equation above could be turned into this form by expressing the fraction as partial fractions. When this is done, we find that:

$$I(s)=\frac{0.003258}{s+1382}-\frac{0.001258}{s+3618}$$

(To confirm this working, add the two partial fractions and verify that you get back to the fraction of **Eq. 134**). A search through the table of Box 1 shows that expressions of this form are transforms of the decay function, $i(t) = e^{-at}$. We are ready to perform the reverse transform, reading from right to left in Box 1 and substituting appropriate values for a and s:

$$i(t)=0.003258e^{-1382}$$
$$-0.001258e^{-3618}$$

This is precisely the same result as was obtained in Part 13 using the straightforward techniques for solving differential equations. It serves to confirm that the Laplace transform does give the correct result, in spite of the apparently roundabout route from start to finish. The advantage of the Laplace method is that it can be used with differential equations that do not yield readily to the ordinary techniques.

It has taken several paragraphs to work through this example, but the steps in the calculation are few:

1. Transform the differential equation, using Boxes 1 and 2.
2. Simplify the equation to obtain an expression for $I(s)$.
3. Insert initial values and recast the expression so that it consists of terms of the same type as the transforms in Box 1.
4. Find the inverse transforms, using Box 1, to obtain an equation for $i(t)$.

The important point to notice is that at no stage is there any need to differentiate or integrate. The tables of transforms cover almost every case, and tables more extensive than Boxes 1 and 2 are available for the infrequently used functions. The only maths required is simple algebraic manipulation, mostly the finding of partial fractions. This does not seem to appear in GCSE maths syllabuses, so a simple routine for this is outlined in Boxes 3 and 4.

Transformed circuits

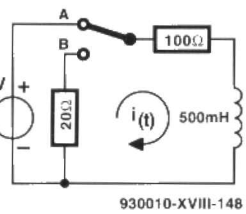


Fig. 148

Figure 148 shows a circuit in which the switch is in position A long enough to reach a steady state. The switch is changed to position B when $t = 0$. At any instant the voltage across the two resistors is $120i(t)$ and, because it depends on the rate of change of current, the voltage across the inductor is $0.5di(t)/dt$. Applying the rules for networks that we first met in Part 4, we can state that by KVL, and with the switch in position B:

$$120i(t)+0.5di(t)/dt = 0$$

[Eq. 135]

This differential equation describes the behaviour of the circuit but, since it does not take into account the initial state of the circuit, we do not yet have an equation for $i(t)$. Because the voltage source is now switched out of the circuit, it is hard to see how we can allow for it. This is where the transform helps out. Transforming **Eq. 135**:

$$120I(s)+0.5\{sI(s)–i(0^+)\} = 0$$

[Eq. 136]

Now we can take the initial current, $i(0^+)$, into account. We find its value by noting that in the steady state with the switch at A, assuming that the resistance of the inductor is negligible, a voltage of 6 V across a resistance of 100 Ω causes a current

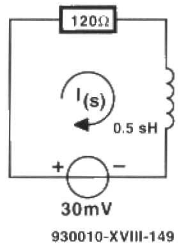


Fig. 149

of $i(0^+) = 6/100 = 0.06$ A. Substituting in **Eq. 136**:

$$120I(s)+0.5sI(s)–0.03 = 0$$

or

$$120I(s)+0.5sI(s) = 0.03$$

[Eq. 137]

Equation 137 is the equation which woulld be obtained by applying the network rules to the circuit of **Fig. 149**. In other words, **Fig. 149** is the circuit of **Fig. 148** transformed into the frequency domain. Note how the inductance depends on s, which has the dimensions of frequency. The circuit includes a voltage source, representing the voltage due to the initial current. From **Eq. 137**:

$$I(s) = 0.03/(0.5s+120)$$

Before we can reverse the transform, we must multiply the numerator and denominator of the fraction by 2 to obtain s in the denominator instead of $0.05s$:

$$i(s) = 0.06/(s+240)$$

Box 1 shows that this is the transform of the decay function. In the transform, a = 240 and:

$$i(t) = 0.06e^{-240t}$$

The graph of this function is given in **Fig. 150**. It shows that the current starts at 0.06 A and decays exponentially, being very close to zero after 20 ms.

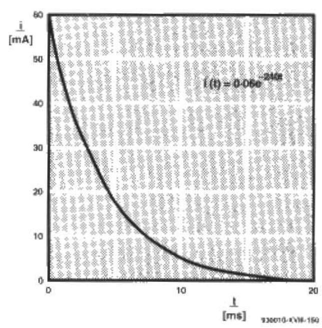


Fig. 150

Next month we look at how to model capacitors and how to calculate initial and final currents and voltages.

Test yourself

1. Use the table of Box 1 to find the transforms of (a) $2e^{-6t}$, and (b) $3\cos4t$.
2. Use the table of Box 1 to find the inverse transform of (a) $3/s(s–3)$, and (b) $28/(s^2+49)$.
3. Write the transform of this equation and solve it, using the starting conditions given: $d^2i(t)/dt^2+di(t)/dt–6=0$, given that $i(0^+)=0$ and $di(0^+)/dt=5$.
4. In the circuit of **Fig. 151**, the input voltage is 10 V at $t = 0$, and ramps down according to the equation $u = 10–200t$. (a) Express the current in this circuit as a differential equation; (b) find the transform of the equation and simplify it; (c) find the inverse transform to obtain an equation which shows how the current varies in time.

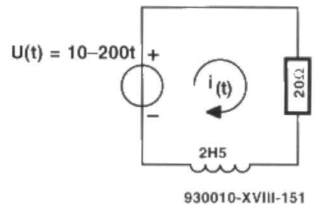


Fig. 151

Answers to Test yourself (Part 17)

1a $\delta z/\delta x = 2+3y$;
$\delta z/\delta y = 2x+3$.

1b $\delta z/\delta x = 3x/2y$;
$\delta z/\delta y = -3x^2/4y^2$.

2. Charge decreases by 0.2 μC, from 198.0 μC to 197.8 μC.

3. $dr/dt = 0.75$;
$du/dt = 100\times10^{-6}$;
$\delta i/\delta u = -1r$;
$\delta i/\delta r = (u–10)/r^2$;
$di/dt = 86.7\times10^{-9}$ when $i = 1$ mA, equivalent to a tempco of 86.7 ppm °C^{-1}.

[930010-XVIII]

RESETTABLE FUSE FOR CARAVANS

Considering the increasing volume of electrical equipment towed along by today's caravanners, people frantically searching for torchlights and then the spare fuses box can be observed almost any night on any camping site. If you have a caravan, and want to keep that unpleasant aspect of holiday making at bay, you are well advised to extend the electrical system with a resettable fuse as described here.

From an idea by E. Bosman

THE true spirit of caravanners and camping enthusiasts seems to have vanished if you look at what some people bring with them to their holiday destination: an electric coffee machine, a hot water boiler, an electric iron, a colour TV sets and even a satellite TV receiver, to mention but a few things. Obviously, the modern holiday maker does not want to forfeit the comfort and luxury he or she has come to value so much at home. At the same time, most of you would also like to think of camping as a 'primitive' pastime: back to nature, idyllic places, elementary cooking, and making coffee on a smoking little fire made from wet wood, that is what it should be about. Unfortunately, as in so many cases, modern technology has changed all that. For better or for worse, that is for you to judge.

In any case, the arrival of more and more electrical equipment in and around the caravan or mobile home greatly increases the risk of fuses blowing at unexpected times. A blown fuse is not a very serious event in itself if it is just the caravan fuse, since that is fairly easy to replace. Provided, of course, you have your torchlight handy when disaster strikes (most of you will know that fuses usually blow when it is dark).

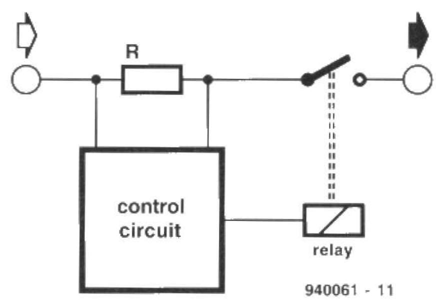
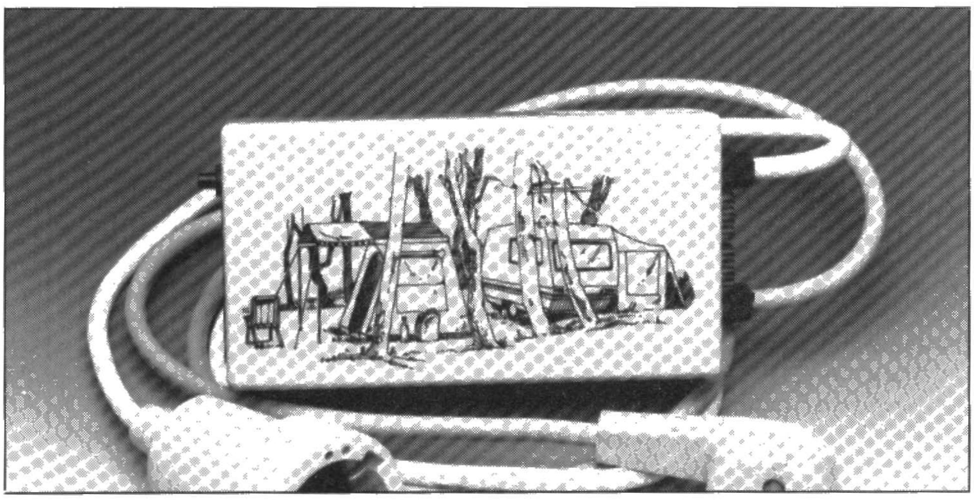


Fig. 1. Basic elements that make up an electronic, resettable, fuse.

Far more trouble, indeed, if the fuse in the caravan connection post on the camping blows. That can happen if it has a lower rating than the caravan fuse. Undesirable as it may be, a blown fuse is often difficult to forestall in the hubbub of getting your caravan parked, connected, and so on. A lot of trouble can be prevented by first asking the camping proprietor about the rating of the fuse installed in the connection post, and install a lighter fuse in your caravan before actually connecting up to the mains network. For example, if the mains outlet is said to have a 10-A fuse, it is best to use a 6-A fuse in the caravan. Undoubtedly the best way to prevent trouble with the mains supply is to insert a resettable, adjustable fuse in series with the (fixed) caravan fuse. The resettable fuse is then adjusted for a trip current just below that of the fixed fuse. Once the resettable fuse is installed, all you have to do is press a button if you are suddenly in the dark as a result of an overload or an short-circuit.

The electronic fuse described here is fairly easy to build, as reliable as can be achieved with simple means, and adjustable for currents between 4-A and 16 A. In other words, a 'must' for all caravanners!



General layout

Although there are several options to designing an electronic alternative to a fuse, there is no way to go round certain basic 'ingredients'. For example, a current sensor is always required to detect the overcurrent level, as well as a device to break the supply. In other words, the design is basically as shown in **Fig. 1**.

A series resistor, *R*, and a relay contact are inserted in the supply line to the load. The voltage drop across *R* depends on the current which flows through the resistor. All that is required in addition is an electronic circuit which monitors the voltage across *R*, and actuate the relay if a certain threshold is exceeded. What remains, is, of course, the question how the electronic control is realized in practice.

Comparators with a memory

To be able to check if a voltage exceeds a certain level, that level needs to be defined as a threshold. Although that can be achieved with the aid of a comparator, it is also desirable for the load

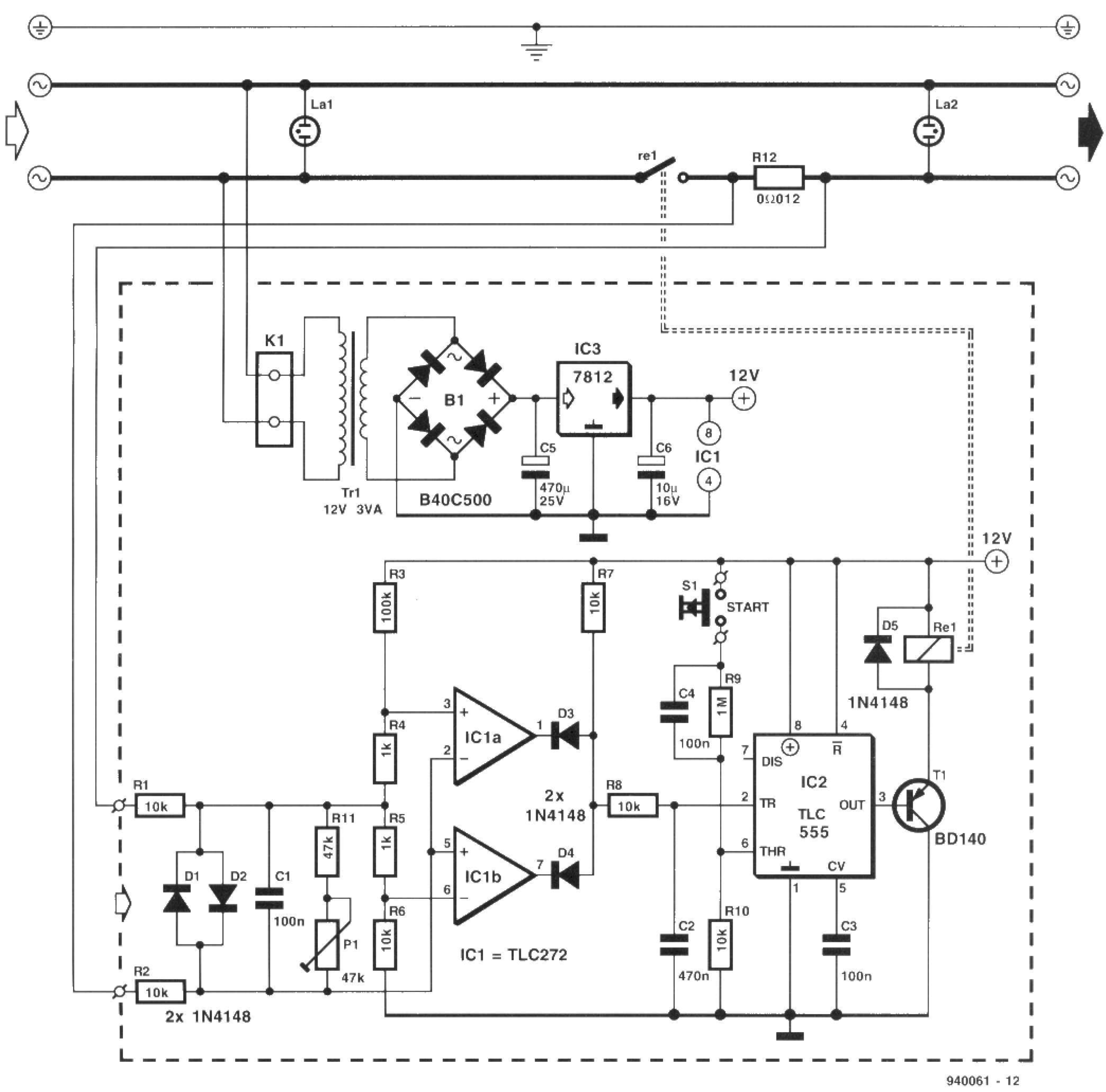


Fig. 2. Circuit diagram of the resettable fuse for caravanners.

voltage to remain switched off (after an overload condition) until the reset button is pressed. In other words, some kind of memory function is in order.

A cursory look at the circuit diagram in **Fig. 2** will inform you that both basic elements mentioned above are present in the circuit. There are even two comparators, so that the circuit responds to the positive as well as the negative half cycle of the mains voltage. Briefly, the operation is as follows. The series ('sensing') resistor is formed by R_{12}. When the voltage across R_{12} rises above +100 mV or −100 mV, the respective comparator IC_{1a} or IC_{1b} toggles, causing bistable IC_2 to be triggered, and to change state also. Consequently, the relay is de-energized, and the bistable holds its state until the 'start' key is pressed.

Let us examine the circuit in more detail. The sensing resistor, R_{12}, has a low value to prevent too much power being wasted by dissipation. Resistors with such a low value are fairly easy to make yourself from (enamelled) copper wire of a known diameter or wire gauge number. Since the value of R_{12} determines the trip level of the electronic fuse, **Table 1** indicates how values of 18 mΩ down to 5 mΩ may be produced for maximum fuse actuation currents of 4 A and 16 A respectively.

Comparators IC_{1a} and IC_{1b} jointly compare the input voltage with a reference voltage derived from the 12-V supply with the aid of R_3-R_6. Of these resistors, R_4 and R_5 drop about 2×100 mV. The alternating voltage dropped by the sensing resistor arrives at the voltage monitor via R_1 and R_2.

I_{max}	R_{sense}	$P_{R(sense)}$	c.s.a.	wire dia.	length
4 A	18 mΩ	0.3 W	0.5 mm^2	0.8 mm (22)	52 cm
6 A	12 mΩ	0.5 W	0.5 mm^2	0.8 mm (22)	35 cm
10 A	8 mΩ	0.7 W	0.5 mm^2	0.8 mm (22)	23 cm
16 A	5 mΩ	1.13 W	0.75 mm^2	1 mm (20)	22 cm

c.s.a. = cross-sectional area.
Nearest SWG value in brackets.

Table 1. Design data to help you make your own shunt resistor.

This voltage is limited by diodes D_1 and D_2, while capacitor C_1 suppresses fast pulses and RF noise. The sensitivity of the fuse is adjusted with preset P_1. The preset acts as a kind of 'fine adjustment' of the fuse value, bearing in mind that the maximum current defined by R_{12} can only be increased, not decreased, by the preset.

As soon as the positive or negative half period of the voltage across R_{12} exceeds the reference level of 100 mV, one of the comparator outputs swings from high to low. Next, bistable IC_2, a

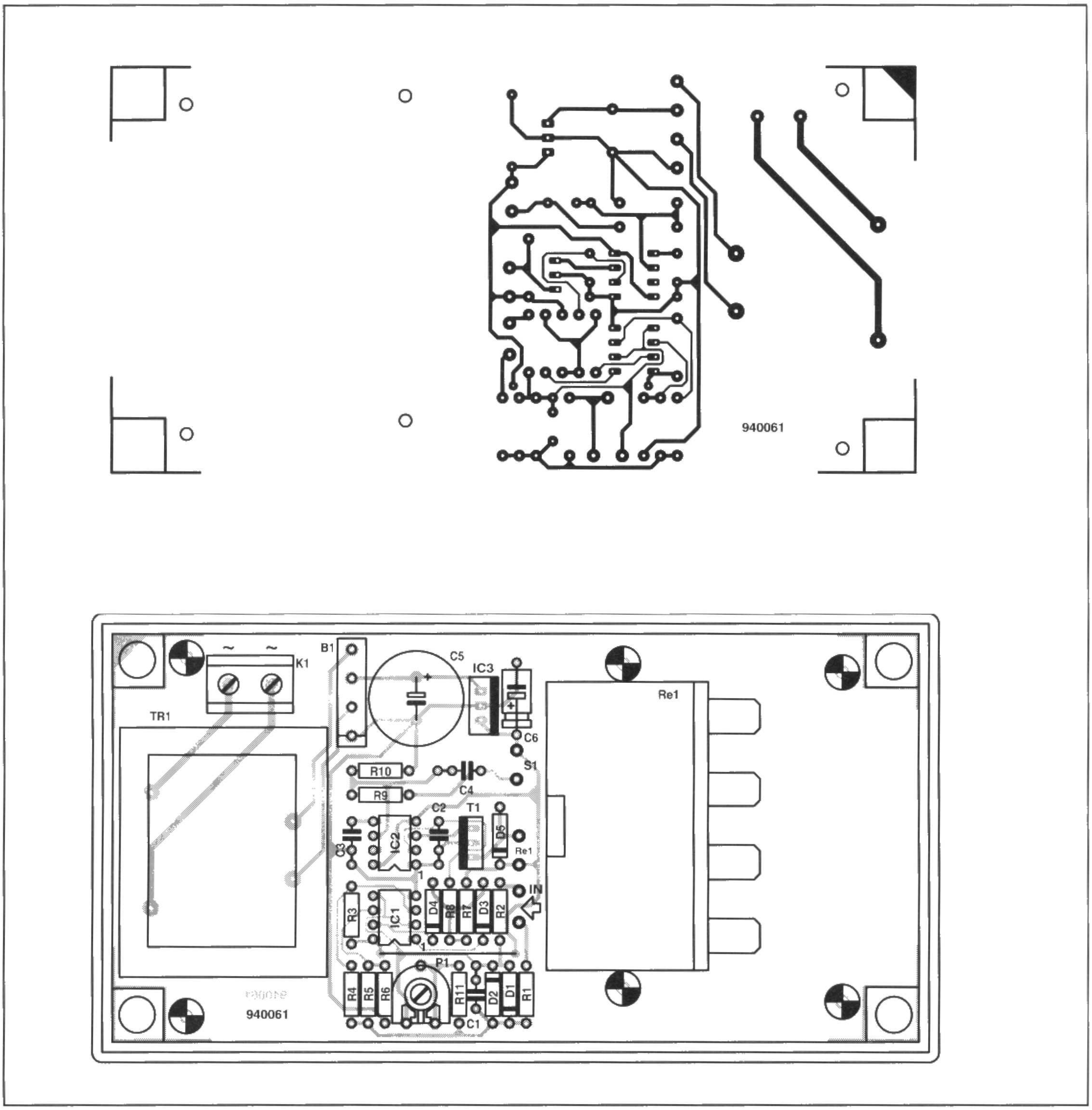


Fig. 3. Track layout (direct reading) and component mounting plan of the printed circuit board designed for the resettable fuse (PCB not available ready-made).

COMPONENTS LIST

Resistors:
R1,R2,R6,R7,R8,R10 = 10kΩ
R3 = 100kΩ
R4,R5 = 1kΩ
R9 = 1MΩ
R11 = 47kΩ
R12 = see table 1
P1 = 47kΩ preset H

Capacitors:
C1,C3,C4 = 100nF
C2 = 470nF
C5 = 470µF 25V
C6 = 10µF 16V

Semiconductors:
D1-D5 = 1N4148
T1 = BD140
IC1 = TLC272
IC2 = TLC555
IC3 = 7812
B1 = B40C500

Miscellaneous:
K1 = PCB terminal block, raster 7.5mm.
S1 = push-to-make button.
RE1 = relay, 12V, w. make contact 240V/16 A (e.g., Conrad Kaco RY-T1L or Siemens V23008 series).
Tr1 = mains transformer, 12V/3VA (e.g., Block VR3112, Monacor VTR3112 or Velleman 1120038M).
La1,La2 = neon light w. series resistor.
Enclosure: dimensions 150x80x55mm (e.g., Bopla E440).

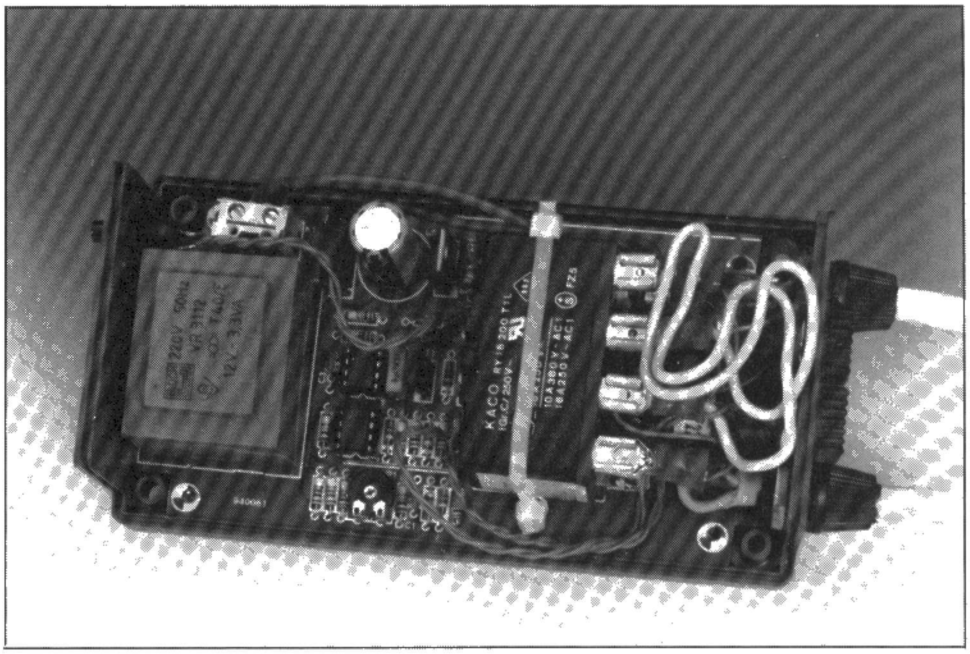


Fig. 4. Completed prototype of the caravan fuse. Note that the current sense wire is insulated and folded. Also note the use of heavy-duty grommets and strain reliefs on the mains in and out cable.

'good old' 555, is triggered via low-pass filter R_8-C_2. The high level at the Q output of the 555 then swings high, causing transistor T_1 to be switched off, and the relay to be de-energized. The 555 now acts as a 'memory', which retains its state until a short pulse is received at pin 6. That happens when the 'start' key is pressed. The relay is then re-energized, and the fuse is 'whole' again, re-establishing the current flow. If, however, the short-circuit or overload condition still exists, the relay will be de-energized again instantly. Keeping S_1 pressed has no effect, since only a pulse resets the 555. In other words, it is necessary to release the key and press it again in any case. Before you do that, however, investigate and clear the cause of the overload.

Short transients in the comparator input signals are suppressed by C_1. Futhermore, the trigger signal for the bistable is purposely delayed a little to prevent the fuse being actuated on every current surge. This is achieved with the aid of network R_8-C_2. The neon lights (with internal resistors), La1 and La2, are added to the circuit to be able to check at any time if the mains voltage is present at the input and output of the fuse. If you wish, you may omit these lights, although we found them very useful on the prototype.

The power supply is entirely conventional, consisting of a mains transformer, a bridge rectifier, a smoothing capacitor and a fixed voltage regulator.

Construction

The artwork of the printed circuit board designed for the resettable fuse is given in **Fig. 3**. Unfortunately this board is not available ready-made through the Readers Services, so you have to have it made, or etch it yourself. Populating the board will present no problems using the component overlay and the parts list.

The main attention during the construction should go to the sensing resistor, R_{12}, the relay, and the way the current-carrying connections are made. **Table 1**, apart from indicating the diameter and length of the copper wire needed for a particular resistance value, also shows the maximum power dissipation. Clearly, the wire will run quite hot at high load currents. The relay is not an 'off-the-shelf' component either, as already indicated by the space reserved for it on the printed circuit board. Most Siemens 'E-card' relays used in *Elektor Electronics* projects are capable of switching up to 8 A. However, the contact rating is reduced to 4 A for continuous use. That may be just sufficient for a 'small' fuse, but is certainly inadequate for the present application, where much higher currents are required. Hence, a much larger relay is used here, the connections of which are formed by clamp or screw terminals. that brings us to the third matter of attention: the electrical connections in the fuse circuit.

At currents of between 10 A and 16 A is it no longer possible, or even allowed (in the interest of safety), to establish electrical connections by soldering. PCB terminal blocks are not suitable either. Normal terminal blocks may be used, although you must be sure to use 16-A types, **not** the smaller (and cheaper) 6-A versions.

By far the best way to connect and wire the circuit is to use car-type 6.3-mm (0.25-in.) wide spade terminals and sockets known as 'AMP' types. These terminals are also used on the relay specified in the parts list. Never use an ordinary pair of pliers to clamp the socket on to the cable — it will yield an unreliable and possibly dangerous connection. Instead, use special 'AMP' type crimp pliers, which also offer a cable stripper and cutter. On the cable sockets, make sure that the hole is properly filled with wire before crimping. If necessary, double the wire.

As a matter of course, great attention should be paid to electrical safety, because the present circuit carries the mains voltage at several points. The enclosure must be a sturdy all-plastic (ABS) type. The prototype was housed in a Bopla Type E440 case, which provides a neat fit for the circuit board

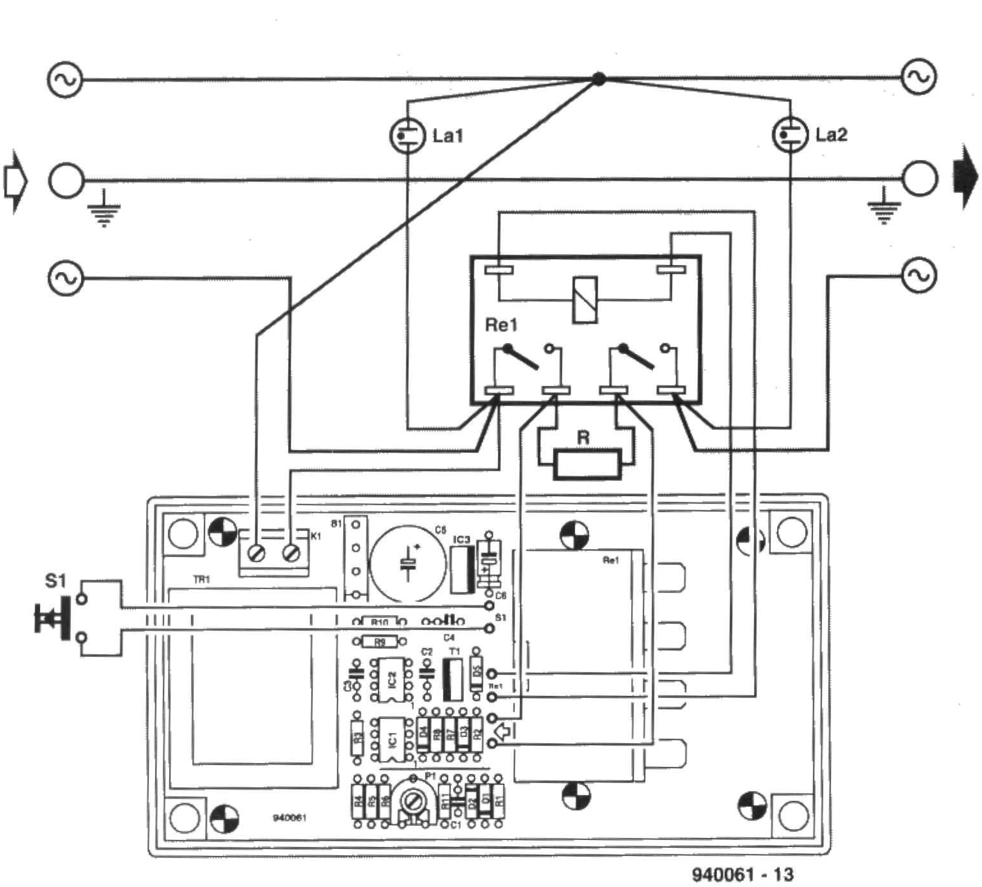


Fig. 5. Wiring diagram. Be sure to use 'amp' type push-on cable sockets and spade terminals to make the current-carrying connections.

and the 16-A relay. The mains input and output cables (220/230/240 V) are three-wire types passed through grommets, and fitted with a strain relief at the inside of the case.

The internal wiring of the fuse is shown in a separate diagram, **Fig. 5**. Do not forget to make a through connection for the protective earth.

Adjustment

The circuit should be functional after a careful check on the wiring and the construction of the printed circuit board. The adjustment of P_1 is not particularly difficult, although it may involve quite a bit of work. A suggested adjustment procedure is given below. In the example, it is assumed that the mains voltage is 240 V.

The preset is set to the centre of its travel, and the fuse is installed. Pressing the start key should close the relay. Both neon lights should then be on. Next, arrange a load which draws a current corresponding to the desired fuse trip level. In case you require a relatively high trip level, say, 12 A, a 2-kW electric heater is good to begin with, since it draws just over 8 A. Add a paint stripper, an electric iron or similar load of about 800 W, and you have a load drawing a current close enough to 12 A.

Turn the wiper of P_1 until the fuse is just actuated (counter-clockwise: less sensitive; clockwise: more sensitive). Be very careful while adjusting the preset, and do not touch any part of the circuit. If the span of P_1 is too small, switch off, and change the length of the sensing resistor as required.

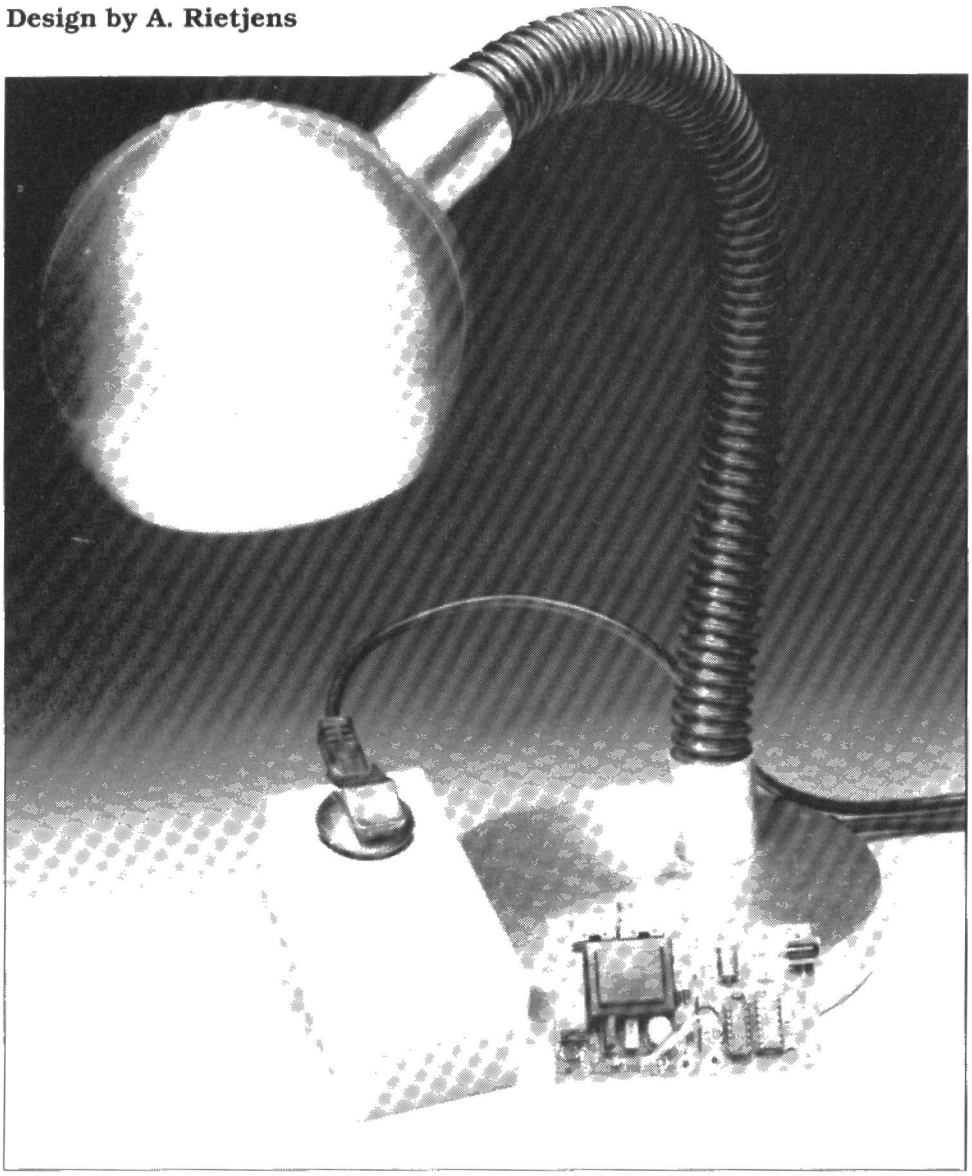
While testing you are will probably notice that the fuse value is a little temperature dependent. The reason is simple. As the circuit heats up, the resistance of the sensing resistor rises, so that the fuse becomes more sensitive. Fortunately, that is no cause for alarm, since the effect occurs with normal fuses also.

(940061X)

INFRA-RED CONTROLLED SWITCH

Infra-red remote control units for TVs and video recorders usually have a large number of press-keys to select a staggering variety of functions. For simple on/off remote control applications, such as opening a door or controlling a lamp, these transmitters are far too complex. That is why a much smaller transmitter and an associated receiver are described here. The system is suitable for switching loads up to 1,000 W using a simple toggle on/off function.

Design by A. Rietjens



MOST infra-red remote controls supplied with today's audio and video equipment use the so-called RC-5 code. The RC-5 standard is based on a set of codes which are used to control the plethora of functions available on modern audio and video equipment.

The pulse codes generated by remote control units are conveyed to the receivers in the audio/video equipment by infra-red light. The transmitter contains an infra-red LED which is switched on and off at a rate of about 36,000 times per second by an oscillator and a switching transistor. In this setup, the oscillator is actuated by code pulses.

The 'packets' of infra-red light transmitted in this way are received on a diode or transistor which is sensitive to infra-red light. Next, the signal is converted back into electrical pulses by a 36-kHz receiver and an associated detector. The pulses are applied to a decoder which is capable of recognizing the transmitted code. Depending on the received (and recognized) code, one of the functions of the audio/video equipment (volume, contrast, program, etc.) is switched.

The encoder contained in the IR transmitter monitors the keypad on the remote control, and converts the code of the key pressed by the user into a corresponding pulse train.

The IR receiver described below uses an unmodulated carrier with a frequency of 32 kHz, which is a simpler signal than the RC-5 code. The infra-red remote control system has a 'toggle' type on/off function, where every key action produces a short 32-kHz burst.

A miniature infra-red transmitter

The circuit diagram, **Fig. 1**, already indicates that the infra-red transmitter is a compact unit.

At the far right in the diagram are two series-connected infra-red LEDs, D_7 and D_8. These LEDs are powered, and start to emit infra-red light, when transistor T_2 conducts.

T_2 receives base current and is switched on when the output of Schmitt trigger NAND gate IC_{4d} (pin 11) goes logic high. Since its two inputs, pins 12 and 13, are interconnected, the NAND gate functions as an inverting buffer.

As already mentioned, the IR LEDs are switched at the rate of a 32-kHz carrier. This carrier is generated by a square-wave oscillator built around IC_{4c}. If pin 9 of IC_{4c} is made high (which happens briefly any time S_1 is pressed), the gate starts to oscillate at a frequency determined by network R_7-C_9-P_1. The preset, P_1, allows the carrier frequency to be 'tuned' accurately to the operating frequency of the integrated IR receiver contained in the receiver (IC_5, see further on).

Since pressing switch S_1 should produce a short infra-red signal rather than a continuous one, the 32-kHz carrier generator built around IC_{4c} has a time limiter consisting of R_5, R_6 and C_8. When S_1 is not pressed, the inputs of IC_{4a} are low because the charge voltage on C_8 has disappeared via resistor R_5. As soon as S_1 is pressed, the transmitter is powered. After a short while, C_8 is charged sufficiently by R_6 to cause the start state of IC_{4a} to be

changed. The voltage at both gate inputs (pins 1 and 2) is then high enough for the output of IC_{4a} to drop from high to low, causing the square-wave oscillator, IC_{4c}, to be switched off. For a new pulse to be generated via S_1, you have to wait a short while for C_8 to be discharged again via R_5.

Since the IR LEDs are switched on and off by short 32-kHz bursts, they do not need the usual current-limiting resistor required for continuous (d.c.) operation. Here, the internal resistance of the battery helps to keep the peak LED current within specifications.

IR receiver

The circuit diagram of the receiver associated with the small IR transmitter is shown in **Fig. 2**. The upper part of the diagram shows the power supply, while the receiver proper is drawn below.

The power supply consists of transformer Tr_1, bridge rectifier D_1-D_4, smoothing capacitor C_3 and voltage regulator IC_3. The latter provides the receiver electronics with a regulated 5-V supply voltage. The relay coil voltage is taken directly from C_3, and appears across the relay coil when transistor T_1 is switched on by the receiver electronics. T_1 then receives base current from pin 9 of IC_{2b}, via resistor R_3.

D-type bistable IC_{2b} is wired as a divide-by-two scaler. As illustrated by the timing diagrams in **Fig. 3**, each positive-going edge at the clock input of IC_{2b} (pin 11) produces a level change at the outputs (pin 8, inverting input; pin 9, non-inverting input). Consequently, the rectangular signal which arrives at the clock input is divided by two before it appears at the bistable outputs. The first clock pulse at pin 11 causes output pin 9 to go high, while the second clock pulse causes it to go low again. In this way, the relay is alternately energized and switched off again, resulting in the previously mentioned 'toggle' function of the remote control system. The apparatus connected to the relay contacts may be a mains operated device. The connection with the relay contacts is made via terminal block K_1.

Selection

The 'on' and 'off' clock pulses which arrive at the clock inputs of IC_{2b} (pin 11) emanate from D-bistable IC_{2a}. Together with monostable IC_{1a}, the bistable forms a selection circuit which serves to ensure that only 32-kHz signals with a certain minimum burst length cause a clock pulse at pin 11 of IC_{2b}. Noise and other spurious pulses are suppressed by the selection circuit

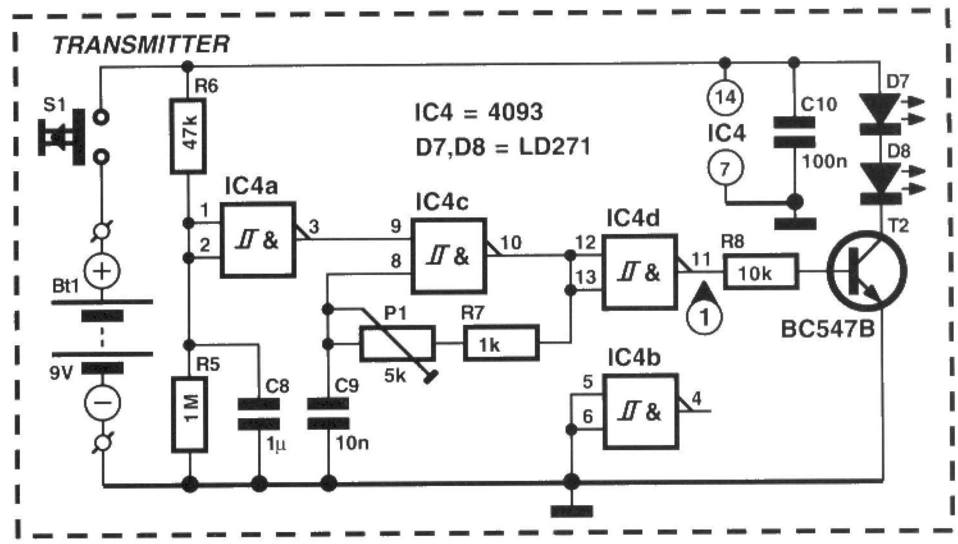


Fig. 1. Circuit diagram of the one-channel on/off infra-red remote control transmitter.

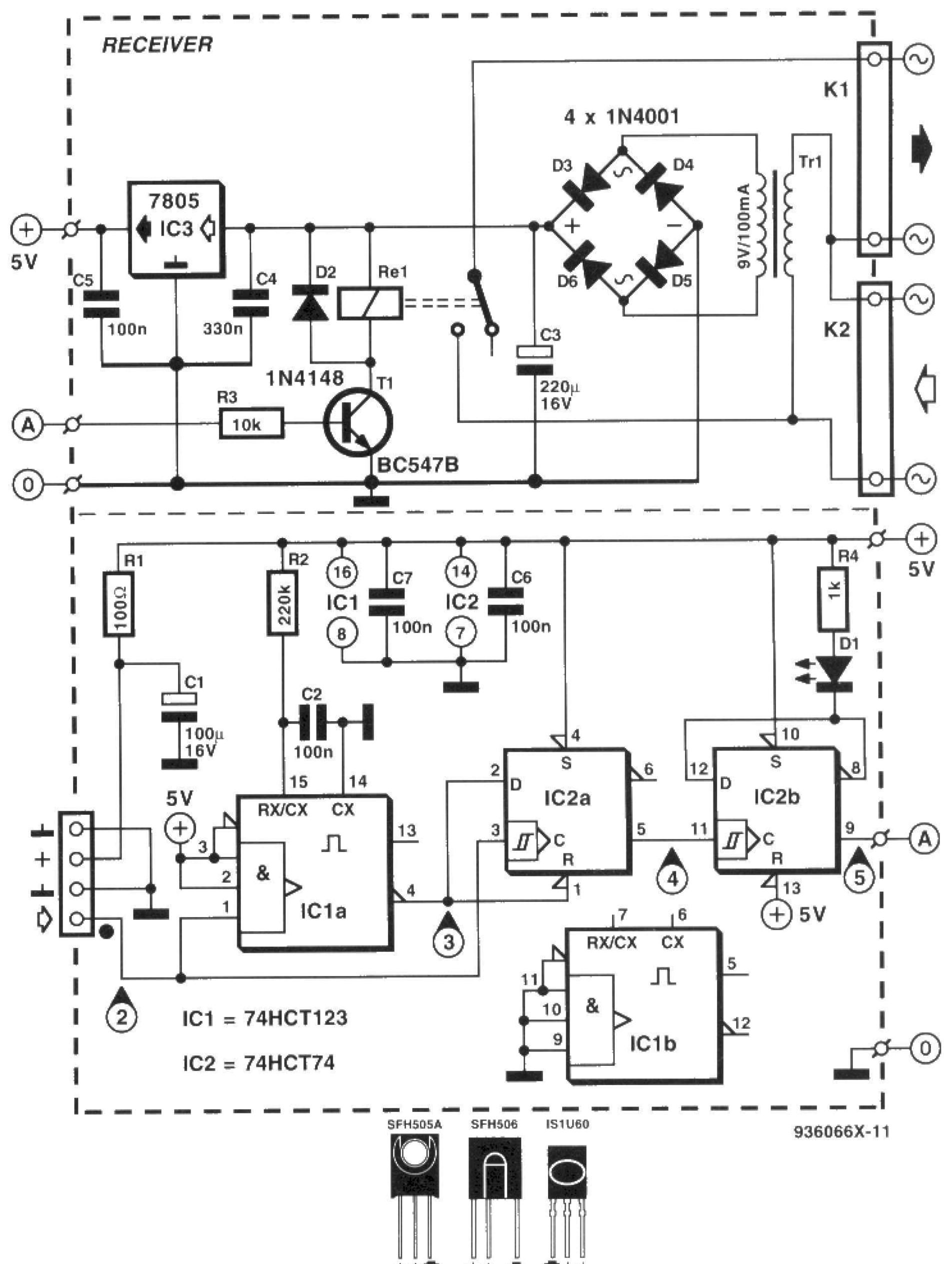


Fig. 2. Circuit diagram of the infra-red receiver, and pinouts of the three IR receiver ICs that may be connected to the input.

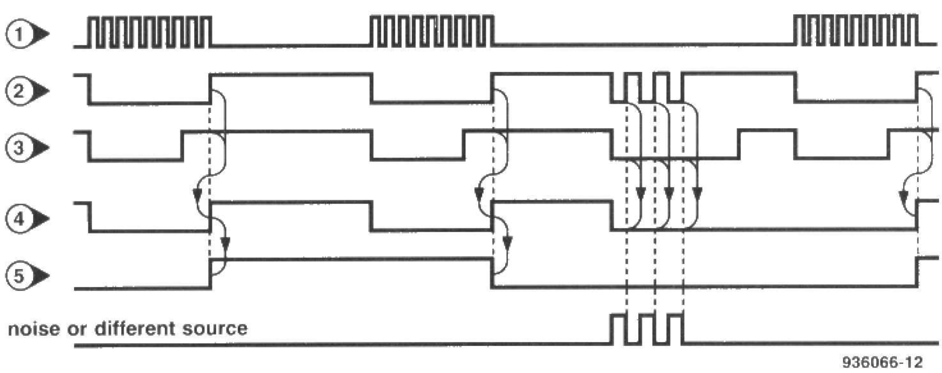


Fig. 3. Timing diagrams of the main signals in the circuit. (1): an IR 'burst' emitted by the remote control, switched on and off at a rate of 32 kHz; (2): the same burst, after detection by IC5; (3): monostable pulse produced by IC1a; (4): an on/off switching pulse; (5): a high level which causes relay Re1 to be energized.

because they are nearly always shorter than the on/off pulses transmitted by the IR remote control.

Figure 3 should help you to understand the operation of the selection circuit. The upper signal, (1), is the 32-kHz carrier. Every time S_1 is pressed, a short burst is generated. The receiver IC, IC_5, contains a complete detector which turns the infra-red light into an electrical signal which is subsequently rectified.

The output of the IR receiver IC is normally high, and swings 'low' only when S_1 on the transmitter is pressed. This 'low' pulse is drawn in Fig. 3 (2). The positive going edges of the pulse serve to control the previously mentioned selection circuit, IC_{1a}-IC_{2a}.

The positive going edges supplied by the IR receiver IC are applied to the

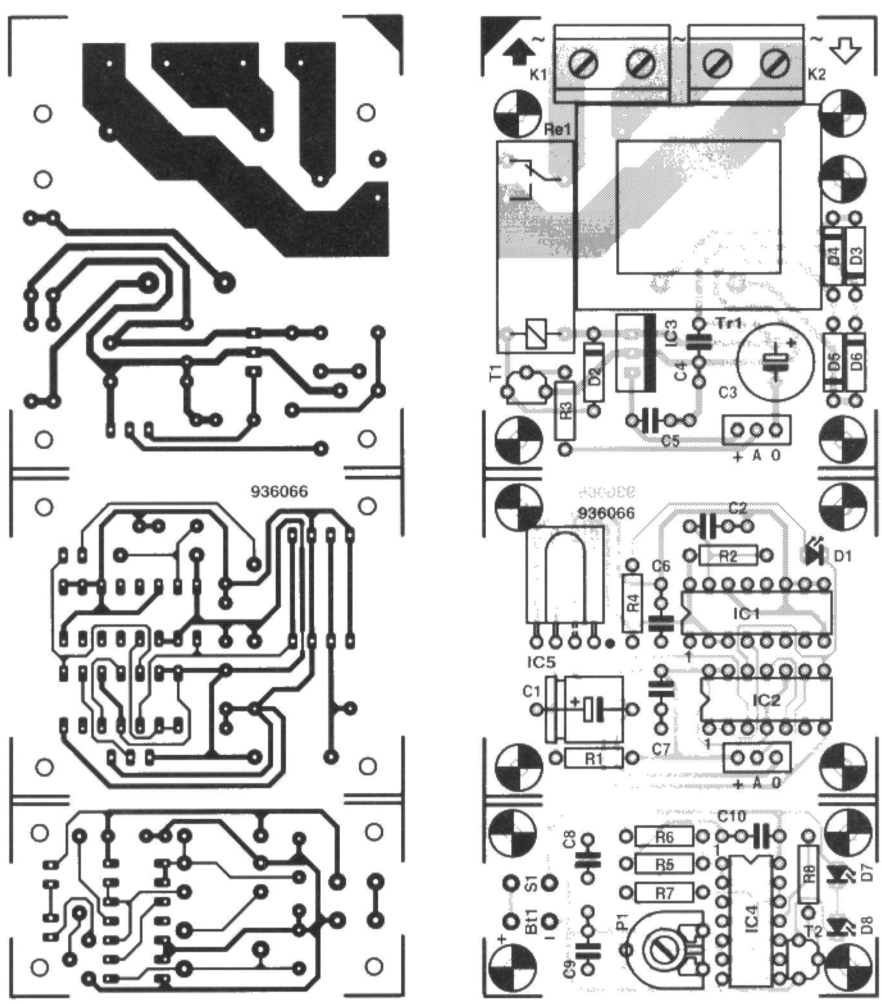


Fig. 4. Track layout (direct reading) and component mounting plan of the printed circuit board for the infra-red controlled switch. This board is available ready-made through our Readers Services (see page 110).

COMPONENTS LIST

Resistors:
R1 = 100Ω
R2 = 220kΩ
R3;R8 = 10kΩ
R4;R7 = 1kΩ
R5 = 1MΩ
R6 = 47kΩ
P1 = 5kΩ preset H

Capacitors:
C1 = 100µF 16V
C2;C5;C6;C7,C10 = 100nF
C3 = 220µF 16V radial
C4 = 330nF
C8 = 1µF
C9 = 10nF

Semiconductors:
D1 = LED 3mm red
D2 = 1N4148
D3-D6 = 1N4001
D7;D8 = LD271
T1;T2 = BC547B
IC1 = 74HCT123
IC2 = 74HCT74
IC3 = 7805
IC4 = 4093
IC5 = SFH505A, SFH506-30 (Siemens) or IS1U60 (Sharp)

Miscellaneous:
K1;K2 = 2-way PCB terminal block, raster 7.5mm.
S1 = push-to-make presskey.
Re1 = 12V card relay, 1 changeover contact.
Bt1 = 9V battery with clip.
Tr1 = 9V/100mA, e.g., VTR1109 (Monacor/Monarch).
Enclosure: e.g. Bopla type SE432DE
Printed circuit board 936066 (see page 110).

input of monostable IC_{1a} as well as to the clock input of D-type bistable IC_{2a}. Every positive going edge supplied by the IR receiver IC causes the monostable to be triggered. The response is a 'low' output pulse at the inverting input, pin 4. The length of that pulse (number 3 in the timing diagram) is determined by R_2 and C_2.

The selection circuit around IC_1 and IC_{2a} will only feed an on/off pulse to pin 11 of IC_{2b} if IC_5 supplies a pulse which is longer than the mono-time of IC_{1a}. In that case, the D-input (pin 2) and the reset input (pin 1) are both logic high when a clock pulse (positive going edge) arrives at the clock input of IC_{2a} (see also Fig. 3). Consequently, the high level at the D input of IC_{2a} is 'copied' to the output of IC_{2a} (pin 9), which goes high. As already discussed, the high level at the output of IC2a causes IC_{2b} and the relay circuit T_1-Re_1 connected to it to be switched on or off.

In the above example, it was assumed that a regular transmitter pulse was received, i.e., one with a length which exceeds that of monostable IC_{1a}. The response of the circuit to interference is illustrated to the right in Fig. 3. Noise (signal 2 in the drawing) is shorter than the monotime of IC_{1a}. The instant the noise is already passed, a clock pulse has been generated at pin 3 of IC_{2b}, although IC_{1a} is still going through its monotime, so that the MMV output, pin 4, is still low. This level keeps IC_{2a} in the 'reset' state, preventing short noise pulses from being clocked through on to the output. Interference is effectively blocked in this way because the individual pulses are shorter than the monotime of IC_{1a}.

Construction

The artwork of the printed circuit board designed for the infra-red transmitter, the receiver and the associated power supply is shown in **Fig. 4**. The printed circuit board is available ready-made through the Readers Services. The three sub-sections (transmitter, receiver and power supply) are separated with the aid of a jigsaw. Cutting is made easy by the dashed lines on the component overlay.

The construction of the transmitter is unlikely to cause problems. Mount IC_4 as the last component, and then do a careful check on your soldering work. To test the transmitter, temporarily connect 'ordinary' LEDs, for instance, red ones, instead of the IR LEDs. The LEDs should light briefly when S_1 is pressed. If this works, mount the IR LEDs, and you can safely assume that the transmitter is functional. As a matter of course, the IR LEDs are fitted in a position that allows them to radiate their invisible light in one, common, direction.

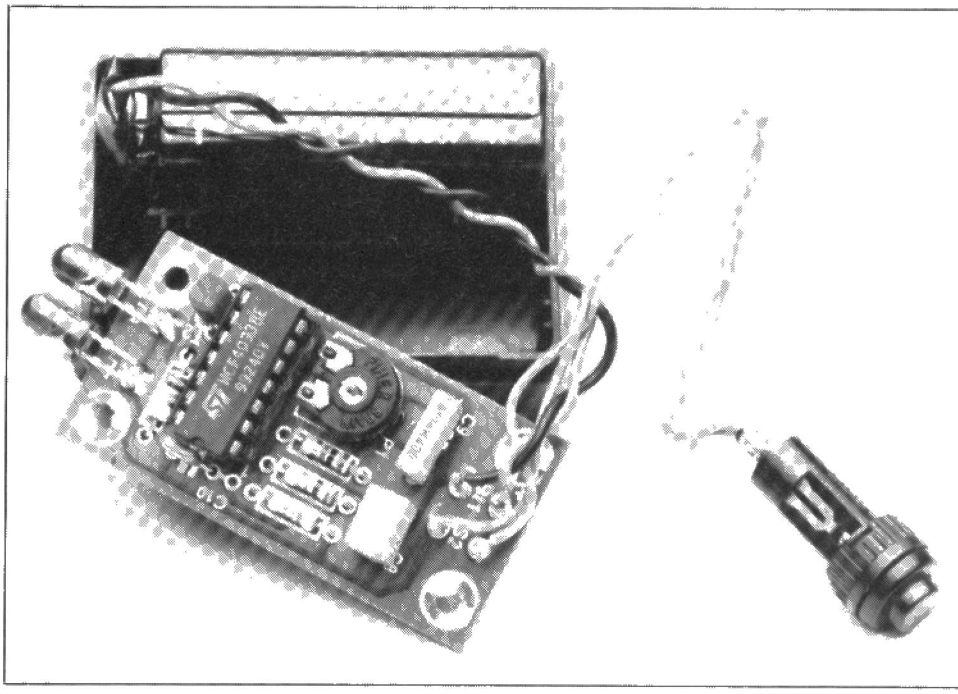


Fig. 5. Prototype of the infra-red transmitter.

Although the transmitter may be fitted in almost any small, plastic case with room for the PCB and the battery, the receiver must be housed in an enclosure which is electrically safe. If you wish, the receiver and the power supply may be fitted in separate cases, when a three-wire cable is used to interconnect these units.

The construction of the receiver is along the same general lines as that of the transmitter, i.e., the ICs should be fitted last. The LED indicator (D_1-R_4) is very useful for testing the receiver. Connect the 5-V supply to the receiver, point the transmitter at the IR receiver, IC_5, and press S_1 repeatedly. LED D_1 should go on and off. Increase the distance between the transmitter and the receiver, and adjust P_1 until no further increase can be achieved.

Since the power supply PCB carries the mains voltage at several points, it must be mounted with great attention paid to electrical safety. All mains carrying wires should be properly isolated using heat-shrink sleeving. Also be sure to use grommets and strain reliefs on the mains cables that enter the case.

The power supply section is the simplest to test: you only have to check if the relay operates (clicks), and D_1 goes on or off, when the transmit key is pressed. If that works, the load to be switched may be connected to terminal block K_1 for a 'final test'. The maximum load power that can be switched is 1,000 W.

(936066)

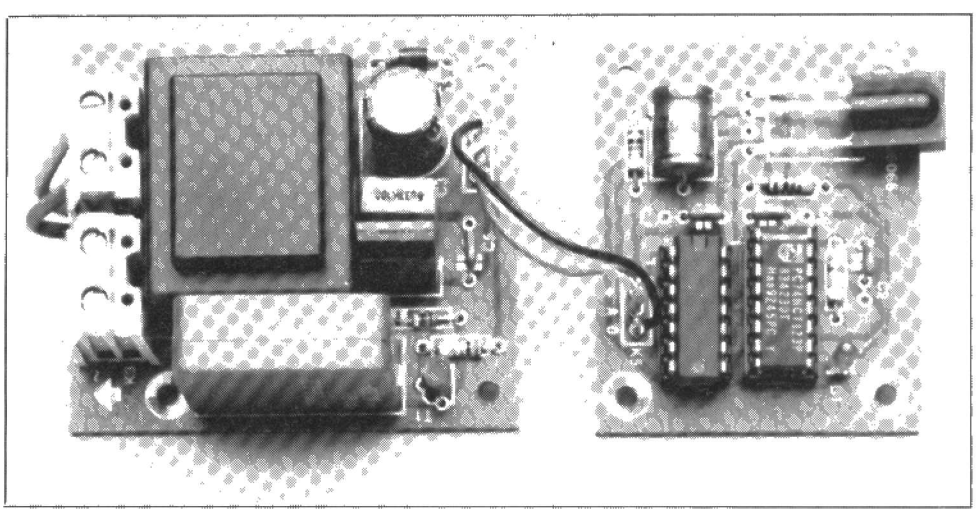


Fig. 6. Fully populated receiver and power supply boards.

PLL-CONTROLLED RAMP GENERATOR

The CD4046 is an integrated phase-locked loop (PLL) with integral voltage-controlled oscillator (VCO) that is used in many digital circuits. In the present circuit, the VCO is used to generate a sawtooth signal. Normally, the VCO produces a triangular signal that is compared with an external digital signal. The VCO in the standard 4046 can be used for digital signals up to 1 MHz; in the HC or HCT version it is usable up to 38 MHz.

In the present circuit, the VCO generates a triangular signal whose rise time is 1000 times longer than its decay time: actually, a sawtooth signal.

Figure 1 shows a section of the internal structure of the 4046, while the complete circuit of the generator is given in **Fig. 2**. The single capacitor normally connected between pins 6 and 7 of the chip is replaced by two capacitors, C_1 and C_2, whose values have a ratio of 1000:1. Capacitor C_2 is discharged rapidly through T_1, which is switched on and off by the signal from the VCO. The capacitor is charged by a FET in the 4046 that functions as a current source.

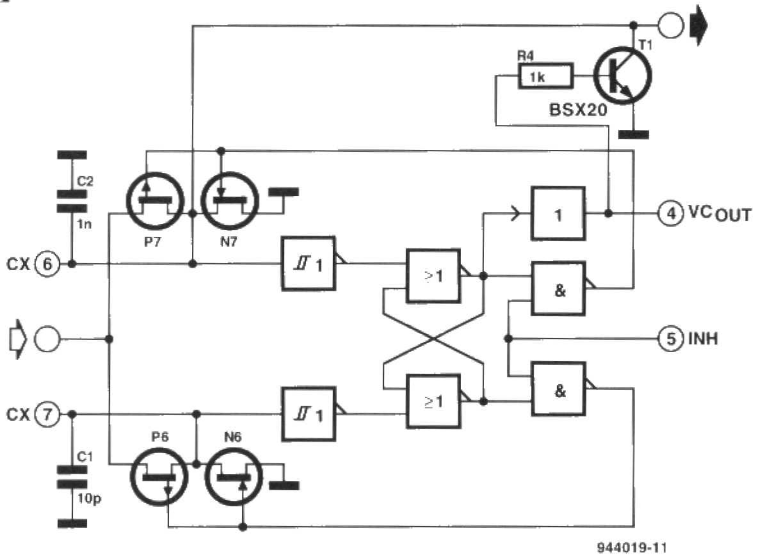
The frequency at which T_1 is switched is, of course, also the frequency of the sawtooth signal. It can be set with P_1, with values as shown, between 20 kHz and 200 kHz.

Since the output is connected directly to the timing section of the circuit, a buffer stage, consisting of a simple FET or operational amplifier, may prove desirable.

The circuit draws a current of about 3 mA.

Design: M.S. Nagaraj
[944019]

1



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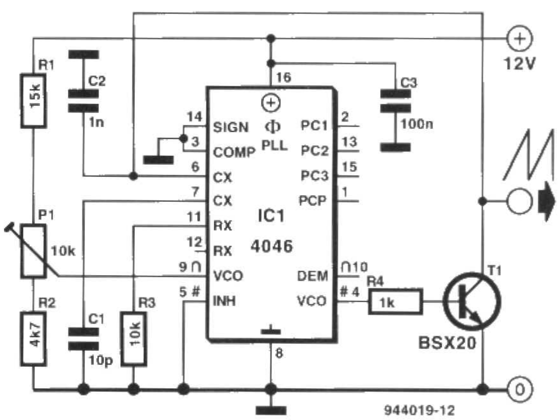
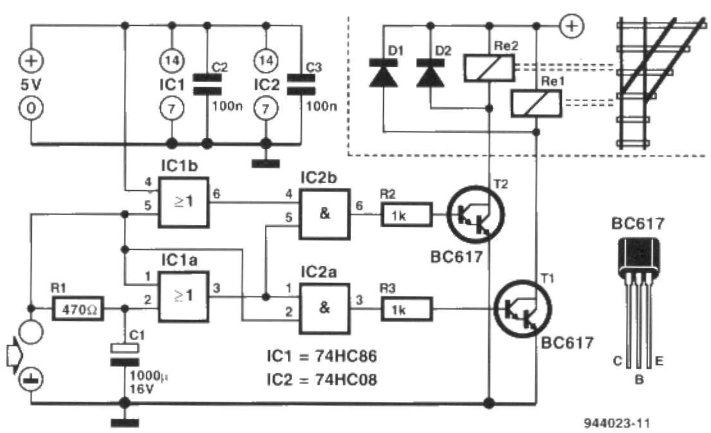


POINTS CONTROL FOR MODEL RAILWAY

Electric points of a model railway system contain two small magnets. To set the points in a given position, one of these magnets must be energized briefly (say, 0.5 s). Any electronic control must, therefore, translate a a change in logic level into a short pulse.

In the present circuit, the digital input signal is applied to XOR gates IC_{1a} (pin 1) and IC_{1b} (which functions as an inverter) and to AND gate IC_{2a}. The input signal is also applied to the other input of IC_{1a} (pin 2) via a delay network, R_1-C_1. As long as the input signal is constant (whether 1 or 0), the levels at the inputs of IC_{1a} are equal and the output, pin 3, is low. The AND gates, IC_{2a} and IC_{2b}, have at least one low level at their inputs, so that their outputs are also low and the transistors, T_1 and T_2, are off.



When the level of the input signal changes, the output of IC_{1a} is high during the time C_1 is being charged. If the input level changes to 1, the output of IC_{2a} becomes high; if it changes to 0, the output of IC_{2b} goes high (because the input is inverted by IC_{1b}).

Depending on which AND gate has a high output, relay Re_1 or Re_2 is energized by T_1 or T_2 respectively. The darlington transistor can switch up to 1 A, which for most points is more than adequate.

It is important that HC types are used for IC_1 and IC_2, because these ensure that the switch-over point is close to half the supply voltage. If standard types are used, T_1 will conduct for much longer than T_2.

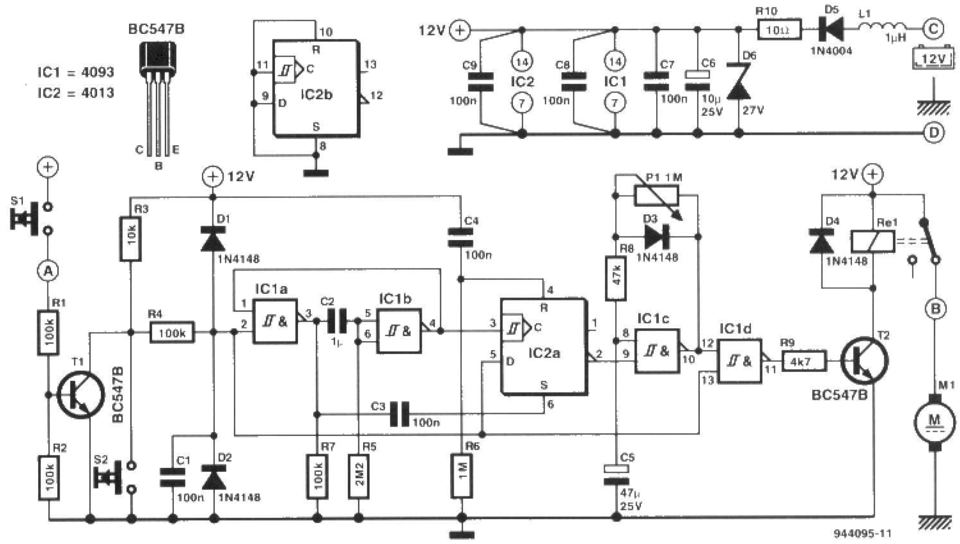
The supply voltage for the points relays should be about 15 V.

Design: M. Averkvist
[944023]

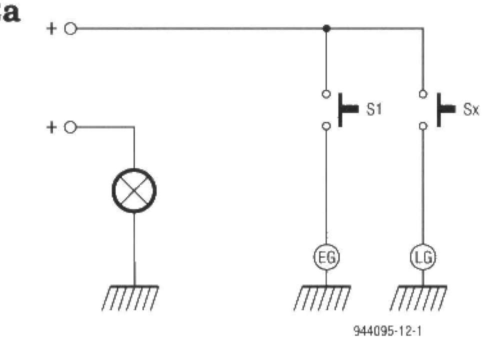
REAR WIPER INTERVAL SWITCH

Many cars have a rear wiper that is controlled by a switch with a make-break combination. In modern cars, the wiper is operated simply by pressing a push-button (with a make contact). In such a case, there is an electronic circuit between the switch and wiper motor, to which an interval circuit can be added. The present circuit allows the time interval to be set between 2 s and 22 s with a potmeter. Operation of the wiper remains via the push-button. If this button is pressed briefly, the wiper travels once and then stops; if the button is held down for more than 2s, the interval circuit comes into operation. Pressing the button briefly again switches off the interval circuit. The advantage of the design is that the existing push button is used: it needs no additional switches as many other designs do.

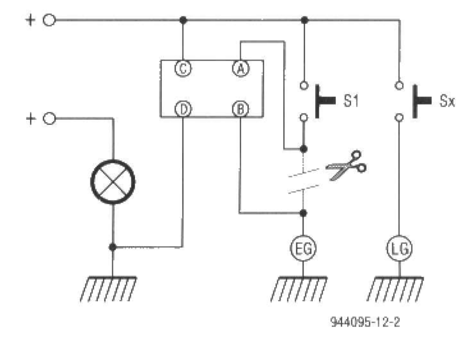
1



2a



2b



In the diagram of **Fig. 1**, S_1 is the existing push-button. When this is pressed, T_1, which functions as an inverter, is on. The level at pin 1 of IC_{1a} is then pulled low via debounce network R_4-C_1-D_1-D_2, whereupon monostable multivibrator (MMV) IC_{1a}-IC_{1b} is started. Its mono time is determined by R_5-C_2, which here is about 2 s.

At the same time that the MMV is started, D-type bistable IC_{2a} is set via C_3, whereupon pin 2 goes low. This results in the output of IC_{1c} going high: the relay is then energized via IC_{1d} and T_2 for as long as S_1 is pressed.

After the mono time has elapsed, the output of IC_{1b} goes high again and the bistable receives a clock pulse. At that instant, the status of S_1 is read in via the D-input of IC_{1a}. If S_1 was pressed briefly, a logic is written and pin 2 remains low. This means that the relay can not be energized, so that the wiper makes only sweep.

If S_1 was depressed when the bistable was clocked, pin 2 goes high and astable multivibrator (AMV) IC_{1c} is enabled. The relay is energized briefly after every few seconds, depending on the position of P_1, so that the wiper sweeps again and again. That situation persists until S_1 is pressed again. The bistable is then reset and IC_{1c} ceases conducting.

Network R_6-C_4 provides a power-on reset

The power supply has several provisions for keeping interference on the car's electrical system from the present circuit. Inductor L_1 blocks high-frequency interference pulses, while zener diode D_6 ensures that the level of signals filtered out by L_1 does not rise above 27 V.

Figure 2 shows how the circuit is connected in the car (here a Renault Espace). The connection between push-button switch and wiper motor is broken, whereupon the two resulting wires are soldered to points A and B.

If one of the terminals of the push-button switch is connected to ground, it functions as S_2 in the circuit: R_1, R_2 and T_1 can then be omitted. The pole contact of Re_1 is then connected to ground, while the switching contact is linked to the wiper motor.

The only item that has to be fitted on or near the dashboard is the knob of P_1.

Design: J. Seyler
[944095]

SMALL LOOP ANTENNAS FOR MW AM BCB, LF AND VLF RECEPTION

PART 2 (FINAL): PRACTICAL CONSTRUCTION

Last month we examined the basic theory of loop antennas, and demonstrated some of the basic forms of loop antenna, resonating methods and coupling methods. In this second and final installment we will take a look at the actual construction of practical loop antennas, some representative loop preamplifier circuits, and a couple of interesting, if odd, applications of loop antennas.

By Joseph J. Carr, B.Sc., M.S.E.E.

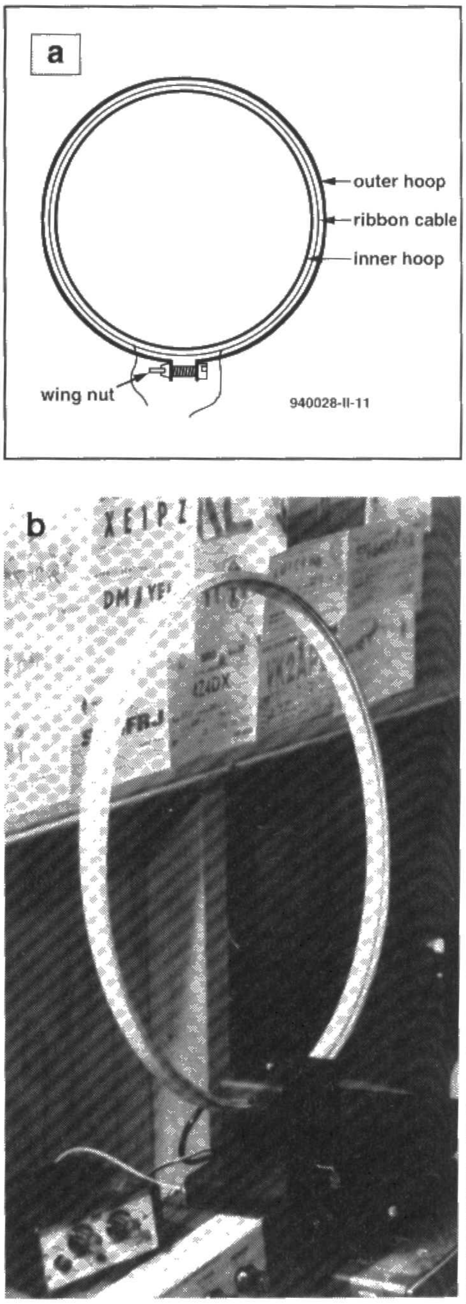


Fig. 13. Embroidery hoop loop antenna: (a) schematic; (b) photo.

Loop construction

Discussing antennas is all well and good, and indeed somewhat intellectually satisfying, but, as they say, 'the devil is in the details'. Unless the loop antenna is to remain a theoretical construct you read about in this magazine, it must be somehow rendered into practical form. And that is where the going can get a might sticky.

Hoop loops

Perhaps the stickiest form of loop to make is the low frequency circular loop. At higher frequencies, a single length of RG52/U coaxial cable can be formed into a satisfactory single-turn loop antenna. Indeed, the amateur radio literature contains numerous examples of 14 through 54 MHz amateur band portable 'fox hunting' RDF loop antennas made from coaxial cable. At lower frequencies, however, the problem becomes a bit more difficult, although not impossible as you will soon see.

A solution that I found is shown in **Fig. 13a** and **Fig. 13b**. A schematic of the hoop loop is shown in Fig. 13a, while the actual antenna that I built is shown in Fig. 13b sitting at my ham radio station. A black plastic box contains a preamplifier and resonating capacitor (box should have been metal, by the way).

The core of this loop construction idea came to me while visiting a crafts store frequented by my wife: it is an embroidery hoop. These products are two-piece circular wooden or plastic hoops sized such that one fits inside the other. The larger outer hoop is broken at one point, and fitted there with a thumbscrew assembly for tightening the outer loop against the inner loop in order to hold fast the fabric the embroiderist is working on.

The first version that I built used multi-conductor ribbon cable with each wire cross-connected to its adjacent mate (more later). In Fig. 13b you can see the outer edge of the ribbon cable protruding beyond the width of the wooden hoop.

Picture frame loop

Another approach is the planar wound loop of **Fig. 14**. This loop antenna was built with supplies from the same crafts store, but this time wooden picture frame material was used. The frame material is intended for do-it-yourself framers, and comes in 2-foot and 3-foot lengths. Each length is cut with tongue and groove, and slanted 45 degrees at each end, so that they joined to form either a longer straight section or a right angle joint. I used four 2-foot sections to form the square loop shown in Fig. 14.

The winding of the loop antenna in Fig. 14 is ribbon cable, as before, but in this case it is thumb tacked to the wooden frame. Care must be taken to push the thumb tacks through the ribbon cable between conductors in to not harm them. I could discern no effect on the performance of the loop antenna from the thumbtacks passing through the cable. Note that the cable is folded over on itself at each corner in order to make the turn. This also had little or no noticeable effect on the performance, although I should imagine that performance deteriorates at least somewhat.

Ribbon cable windings

The previous two antennas, and one to be shown shortly, use computer ribbon cable as the antenna windings. The idea is to cross-connect wires in order to form

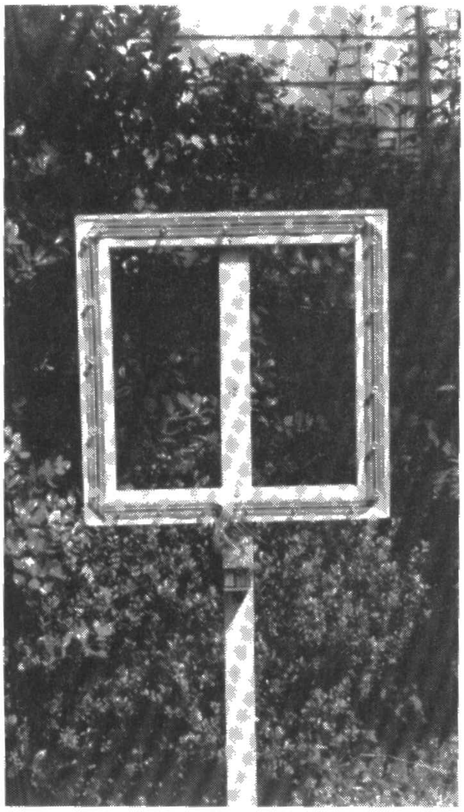


Fig. 14. Picture frame loop antenna.

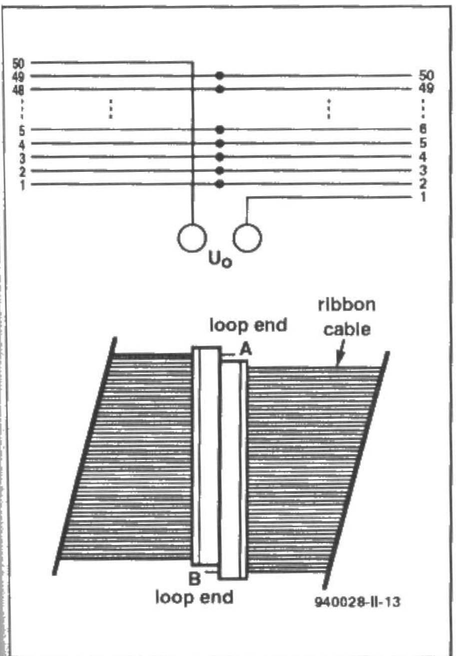


Fig. 15. Cross-connection of ribbon cable conductors to form a continuous loop.

a continuous loop. In the 50-conductor version shown in **Fig. 15**, the connections to the loop at wires 1 and 50. At the beginning of the loop, the other end of wire no. 1 is soldered to wire no. 2, the other end of no. 2 to no. 3 and so on until one end of no. 49 is soldered to one end of no. 50, ... with the remaining free end of no. 50 becoming the connection for the loop.

For small loops (under 1 metre squared), which use only a few turns, one might want to purchase the pre-cut type of ribbon cable that has a single-row in-line connector on each end (female on one end and male on the other). The connectors can be wrapped around the loop frame and then fastened together one pin off (see inset to Fig. 15). The loop end at 'A' is a normal pin from the male connector, while that at 'B' is a loose pin or wire inserted into the opposite end of the female connector.

Cross loop

Figure 16 shows several aspects of the traditional cross loop. I made several of these loops from spruce wood purchased from hobby shops that cater to model builders. The spruce is typically sold in the same display as balsa wood stock. The stock that I purchased from an American source was 24 inches (60 cm) by 3 inches (7.6 cm), and was 3/16 inch (4.8 mm) thick. Two lengths were needed. Each length was notched at the middle ('c') half the width, so that the two pieces could be fit together to form a cross as seen in **Fig. 16b**. At the ends of the wood pieces small slits, the width of a jeweller's or jigsaw blade, were cut to a depth of 6 mm. These slits hold the #26 (0.45-mm dia.) enamelled wire that form the main loop. Holes 0.042-inch (1-mm) in diameter are drilled 12 mm from each end, in the centre of the piece. These holes are for the single turn of the coupling loop.

Figure 16b shows the basic assembled structure of the cross loop. At the junction of the two pieces a set of four 1-cm square stiffeners are glued into place. For better strength, a small screw passing through stiffeners on opposite sides of the same wooden member might be in order.

Figure 16c shows the finished loop with the wires strung. The support for the loop (besides the dead pine tree beside my house) is a 2.5-cm wooden dowel about 1.5 metres in length. This antenna proved quite useful on 75/80 metre ham

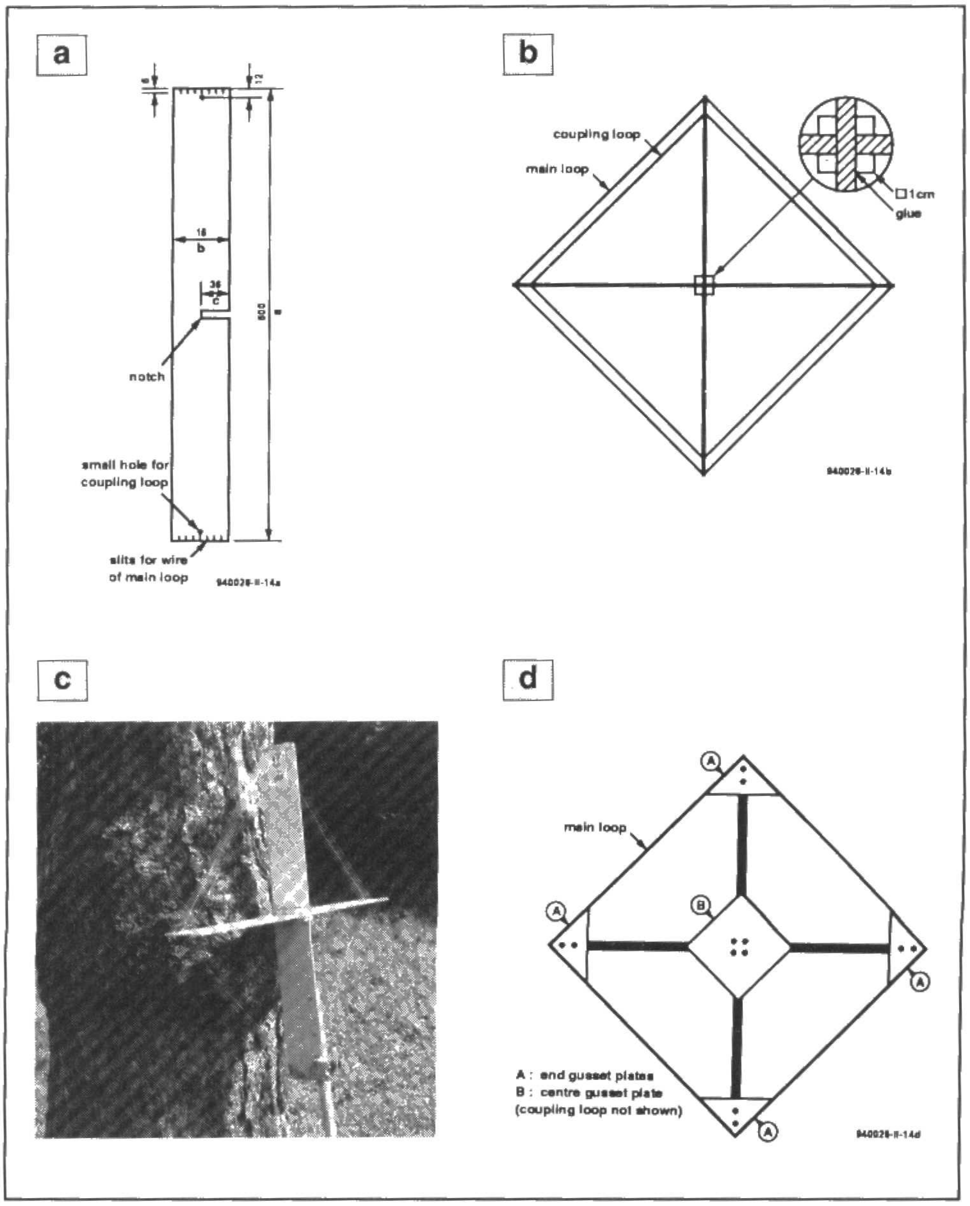


Fig. 16. Cross loop antenna: (a) form of each element; (b) assembly; (c) photo; (d) improved form.

bands, and could be used for fox hunting applications.

Subsequent use of this antenna proved the design to be mechanically weaker than I prefer. In order to overcome this defect, I added the corner gussets and a centre plate gusset, as shown in **Fig. 16d**. The gussets were cut with 45-degree angles from the same type of stock as formed the cross pieces. Additional 1-cm pieces can be placed behind the corner gussets to improve stability, if necessary.

Large box loop

At VLF frequencies a large box loop is sometimes in order. Loops with dimensions of 1.5 to 3 metres squared are found in the literature. Large square loops are somewhat more difficult to build because mechanical stability becomes a larger issue, especially when the loop is installed outdoors. When wind is a factor, the 'sail area' of the loop becomes a serious issue. In this section we will take a look at a large box loop made with substantial materials (**Fig. 17**).

The basic design of the loop is a square frame stabilized by corner gussets, as shown in Fig. 17. The sectioned view is shown in the inset. The sides of the elements ('A') are made from corner moulding of the sort sold to homeowners at do-it-yourself lumber stores. Use 0.625-inch (1.6-cm) to 1-inch (2.54-cm) stock. The stock is glued and screwed to the backplate ('B'), forming a U-shaped channel. The backplate can be anything from 0.25 (6 mm) to 0.625 inch (1.6 cm) thick, and as wide as needed to accommodate the moulding and the wires ('D'). The wire is laid into the channel, and can be either wound enamelled wire or ribbon cable. In the case of the ribbon cable, for VLF operation, two or more layers of ribbon cable can be wound over top one another (although this approach can seriously increase stray capacitance).

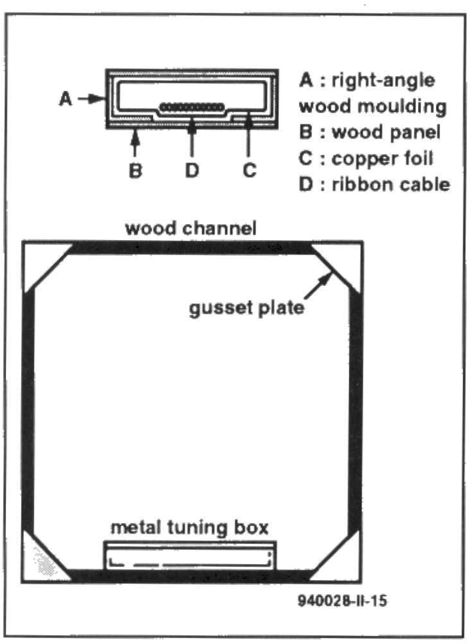


Fig. 17. Form of large loop antenna.

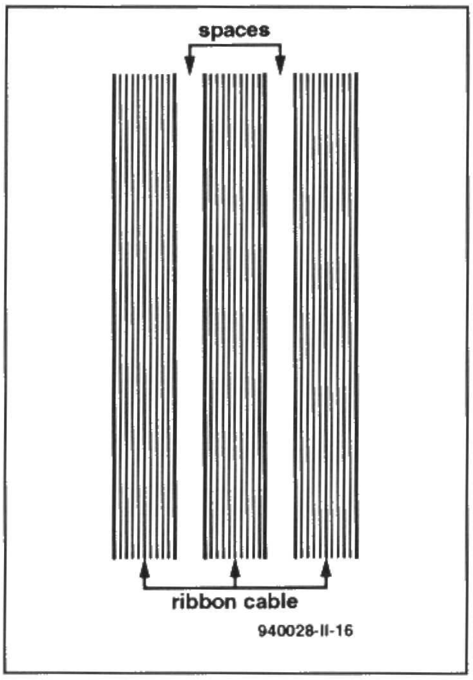


Fig. 18. Low capacitance winding separates groups of conductors.

If the loop antenna is to be shielded (a good idea), then line the U-shaped channel with copper foil ('C') prior to installing the wire. Once the wire ('D') is installed, inspected and tested, then the free foil ends can be folded over on itself and soldered together, completing the shield. Keep in mind to leave a 1-2 cm gap in the foil shield opposite the feedpoint of the loop.

Once the loop is completed, cement a cover ('E') over the U-channel. This cover can be of the same stock as the backplate ('B'), although thinner stock would also suffice. The cover plate can be slotted at the feedpoint in order to bring the cable into the tuning box, where all connections are made and both the tuning capacitor and preamplifier (if either are used) are located.

Reducing stray capacitance

All coils have a certain amount of unwanted capacitance along with the normal inductance. This stray capacitance makes the coil self-resonant at some frequency that is hopefully far higher than the normal operating frequencies. Loop antennas are no exception, and can exhibit rather large stray capacitance numbers.

The stray capacitance does not normally bother loop constructors, and indeed may help. For example, when the variable tuning capacitor (400 pF or so) is insufficient to resonate the loop at some desired frequency. The stray capacitance of the loop may well permit use of smaller add-on capacitors to achieve resonance. But at other times, such as when the self-resonant point is forced too low, the stray capacitance has a bad effect on operation of the loop.

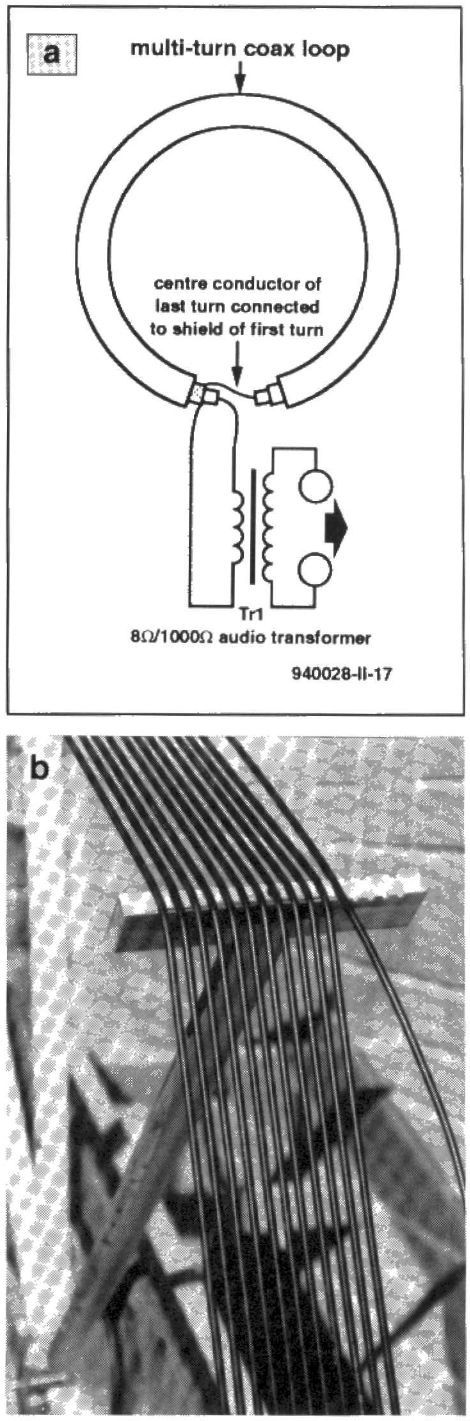


Fig. 19. Coaxial cable 100 kHz loop: (a) schematic. Note that the centre conductor of the far end is connected to the shield at the feedpoint; (b) close up.

Conversations with some radiosolar observers (Taylor and Stokes 1992), while researching another article, showed me that stray capacitance was a big problem for them ... and one which they had overcome. A typical 20 to 30-kHz VLF loop used by radiosolar observers to detect sudden ionospheric disturbances (SIDs) are square, 1 to 2 metres on a side, and wound with 100 to 150 turns of wire. They use an arrangement similar to **Fig. 18** to reduce the stray capacitance effect. Whether ribbon cable or free winding is used, the

windings are separated into groups of 20 to 50 turns, and each group is spaced about 2 cm from the adjacent groups. All of the groups are connected in series with each other in order to form a single continuous loop antenna.

Coaxial cable 100 KHz loop

A reader wrote to me and provided the design of **Fig. 19** (Ingram 1993). The original antenna was designed to receive LORAN-C navigation signals in the vicinity of 100 kHz. The antenna element consists of 16 turns of RG-59/U 75-Ω coaxial cable (in Fig. 19a only one turn is shown), on an average diameter of 2 metres, connected such that the centre conductor of the last turn is soldered to the outer shield of the first turn at the feedpoint. The introductory photograph with Part 1 and the drawing in **Fig. 19b** show the mechanical structure of this antenna. Coupling is provided through an 8:1000 Ω audio transformer. Although a bit large, the coaxial cable antenna should provide very good, low-noise reception because of the Faraday shield manner of the construction. The designer claimed that a 500-kW 100-kHz station at a distance of nearly 500 km produced a signal of 1,000 μV into the 50-Ω input of the receiver.

Coupled ferri-loop

A loop antenna that is a modification of one of Marris' designs (Marris 1992) is shown in **Fig. 20**. The circuit is shown in Fig. 20a and the actual antenna is in Fig.20b. This antenna is made by embedding 7.5-inch (19-cm) ferrite rods in 10-inch (25.4-cm) lengths of PVC plumbing pipe. Each ferrite rod is wrapped with electrical or masking tape to support it when it is force-fit ('gently') into the pipe. I found that a 1-cm diameter rod, when inserted into a 2.5-cm o.d. pipe, required about 14 turns of **3-M** brand black electrical tape to hold it firm when pressed into the pipe.

The windings consist of whatever number of turns are required for operation at the desired frequency. In an antenna meant to work in the 2 to 5-MHz region, including 75/80-metres per Marris' design, I used ten turns of wire, and the ferrite rods were the μ=800 type. Lower frequencies would require higher number of turns, and possibly the μ=1,200 or μ=2,000 ferrite rods.

Three of the four sides of the ferri-loop are identical to each other, and are similar to Marris' design. The sides are held together with cemented 90-degree PVC pipe elbows. The fourth side, however, differs from the other three. It is fitted with a tee-connector. The winding on this side is split into two halves of five turns each. Like the other windings, these are exterior to the PVC pipe. The coupling winding (L_5) consists of 5 turns of wire wound directly on the ferrite rod that forms L_{1A} and L_{1B}. The connections to L_5

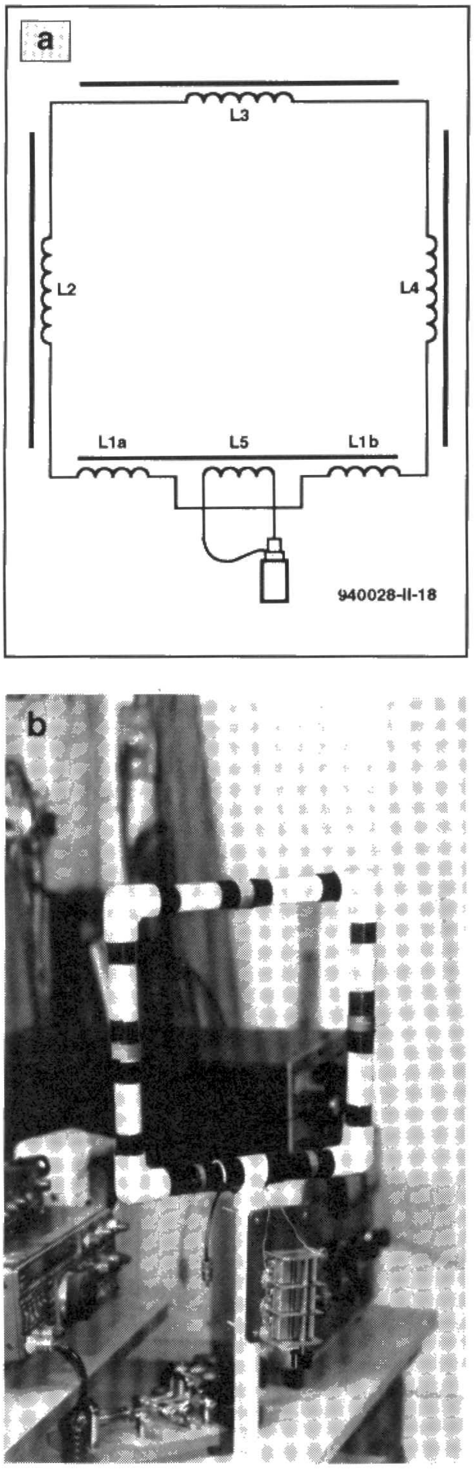


Fig. 20. Ferriform antenna: (a) schematic; (b) photo.

are connected to very thin shielded wire or coaxial cable, and routed to the receiver or preamplifier.

Loop preamplifier circuits

The signal levels obtained by loops is quite low, even when the loop is tuned to the received frequency. As a result, it is common practice to boost loop output using a preamplifier stage. Although almost any preamplifier will suffice, if it cover the desired frequency range, there are several designs which seem most popular, and these will be discussed below.

Whichever amplifier is selected, it must be capable of amplifying the range of frequencies covered by the loop. The amplifier that might be right for the VLF amplifier may or may not also be suitable for operation in the LF or MW portion of the spectrum. Several devices present themselves quite well in this respect, however. For example, the **Mini-Circuits** MAR-x series of MMIC chips may be designed for VHF through low-microwave applications, but they also work well at VLF through HF as well. Similarly, some specialist integrated circuits houses, such as **Burr-Brown**, offer operational amplifiers and operational transconductance amplifiers with gain-bandwidth products of 150, 200, 350 and 500 MHz. These amplifiers can easily be used at VLF through MW frequencies. Another **Burr-Brown** product is the 35 MHz VCA-610 voltage controlled amplifier. This device features a high impedance input, a low impedance output, and is voltage controllable over a range of ±40 gain for a control voltage range of ±2 V. The VCA-610 was designed for ultrasonic medical imaging applications at frequencies similar to those used by VLF through MW radio stations.

As appealing as the above approaches might be, however, except for the MAR-x chips these solutions are also a bit expensive for hobbyist applications. So let us now turn our attention to some circuits using easy-to-obtain components that are low in cost.

Figure 21 shows the circuit of a loop antenna preamplifier that can be built

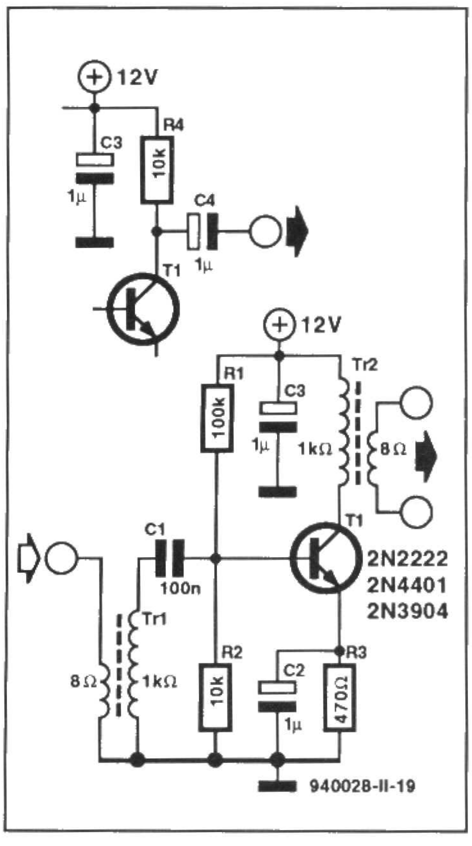


Fig. 21. Single-ended loop preamplifier.

with ordinary hobbyist grade components, yet functions at frequencies up to 70 kHz or so. This circuit uses an ordinary garden variety NPN silicon transistor for amplification. I have used 2N2222, 2N3904 and 2N4401 (latter preferred) for this application, all successfully. The biasing is the ordinary resistance voltage divider (R_1-R_2) with emitter stabilization (R_3).

The capacitor across resistor R_3 is set to have a reactance at the lowest operating frequency of <0.1R_3, or 47 Ω. The idea behind setting the value of C_2 is to keep the a.c. path to ground as low a reactance as possible, while maintaining the d.c. level caused by the current flow in R_3. In practice, this goal is very easy to achieve. At 10 kHz, for example, a 1-µF capacitor has a reactance less than 16 Ω, so falls well within the R/10 rule of thumb. Using a 2.2 µF or 3.3 µF comes closer to a more ideal R/100 rule.

For VLF receivers in the <100-kHz range, almost any electrolytic capacitor will suffice for C_2, although tantalums are preferred. At higher frequencies, however, other forms of capacitor are in order. The **Panasonic** V-series capacitors are available in values up to 1 µF, but are not electrolytics (and thus are not polarity sensitive). These capacitors make a better choice for higher frequency units. Also, some builders actually parallel a 1 µF tantalum with a 0.1 µF disk ceramic for wider frequency coverage. For the <70 kHz bandwidth of this circuit, however, a single 1 µF tantalum capacitor seems sufficient.

Some people who are located close to AM BCB stations, or other large signal sources, may wish to control the gain of the front-end by deleting C_2 altogether. 'Whistler' hunters, i.e., those who seek to receive natural radio signals (Mideke 1992), often leave the emitter resistor of the first stage unbypassed for exactly this reason. It cuts the gain, but is also cuts the level of the interfering signal.

Transformers Tr_1 and Tr_2 are ordinary transistor radio audio transformers. In the circuit shown here, Tr1 is an audio output transformer used in reverse; i.e., the 8-Ω winding is connected to the loop and the 1,000-Ω winding is connected to the input of the amplifier. In some loop antennas, a 1000:1000 Ω transformer is used to couple the loop to the receiver or preamplifier, and can be used in place of Tr_1.

Both Tr_1 and Tr_2 can be ordinary transistor radio grade transformers for work up to 70 kHz or so. At frequencies to 150 kHz, however, these transformers must be replaced by high grade commercial audio transformers that are guaranteed to be ±1 dB to that frequency. Several such transformers are seen in commercial electronics parts catalogs.

Two output circuit configurations are popular for this preamplifier. One is the transformer coupled version shown in the main circuit, while the other is the resistor-capacitor coupled version shown in the inset. The *RC* coupled version replaces the transformer with a 10-kΩ resistor, and couples the signal to the next stage (or receiver) through a 1-µF capacitor (C_4).

Cascode preamplifier

A cascode two-stage amplifier is shown in **Fig. 22**. This circuit uses a junction field effect transistor (T_1) at the input, and an NPN silicon transistor for the output stage. Common devices such as MPF102 for T1 and 2N4401 for T_2 are sufficient. Transistors T_1 and T_2 are direct-coupled, with d.c. bias applied to T_2 through R_2-R_3.

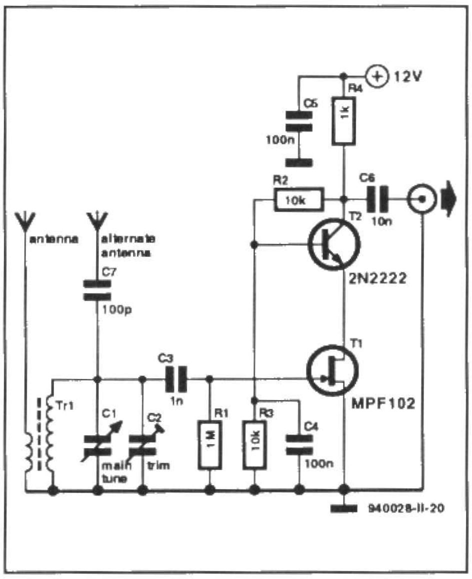


Fig. 22. Cascode loop preamplifier.

The input circuit in this example is tuned to a specific frequency, although in some cases a transformer arrangement such as shown earlier in Fig. 9 might also be used. The inductance needed to tune the loop may be a little hard to come by at the lowest frequencies, in which case two or more coils can be connected in series. In some cases, a xenon tube trigger transformer, such as **Maplin** JE15R, provides a part of the inductance (6 µH), and the rest can be made up with coils in the 10 to 100 µH range. The transformer provides the coupling loop needed to isolate the amplifier from the loop.

Alternatively, the loop itself can be used as the inductor for this circuit. Such an arrangement is not at all uncommon, and works out well, especially when the loop antenna is not located at a remote site from the amplifier (co-location is the usual, and best, practice).

Another alternative is to provide a high impedance antenna input to the preamplifier. If you plan to use a whip, random length wire, or other non-loop antenna, then connecting the antenna to the top of the resonant *LC* tank circuit through a small value disk ceramic or mica capacitor will achieve your purpose.

Push-pull and differential preamplifiers

A lot of loop antenna builders prefer to use push-pull or differential amplifiers for the loop preamplifier job. Any number of possibilities present themselves. For example, an operational amplifier in the differential configuration can be used, if it has a sufficient gain-bandwidth product. Devices such as the CA-3140 (and related chips) or the **Signetics** NE5534 device, are easily available and will work well into the VLF region. Also, devices such as the CA3028, which is popular in amateur radio circles, is also useful for this purpose.

Figure 23 shows a circuit based on common JFET transistors. Each transistor operates in the common source configuration, with the inputs tied to the loop outputs and loop shield. The JFET outputs are combined in a trifilar wound three-winding transformer (Tr_1). The **Mini-Circuits** RF transformers can be used in this application, although at VLF they have a substantial loss (−3 dB or so). One can also wind a variant of Tr_1 using either toroidal or bazooka forms made of low frequency ferrites. The number of turns will depend on the frequency used, and some experimentation is needed. I found that 50 trifilar turns of #30 (0.31 mm dia.) enamelled wire over an FT-68-72 form worked well at 60 kHz in a WWVB receiver, but I have not checked it at frequencies lower than 60 kHz.

Special loop applications

The loop antenna is sufficiently different from other antennas to suggest some in-

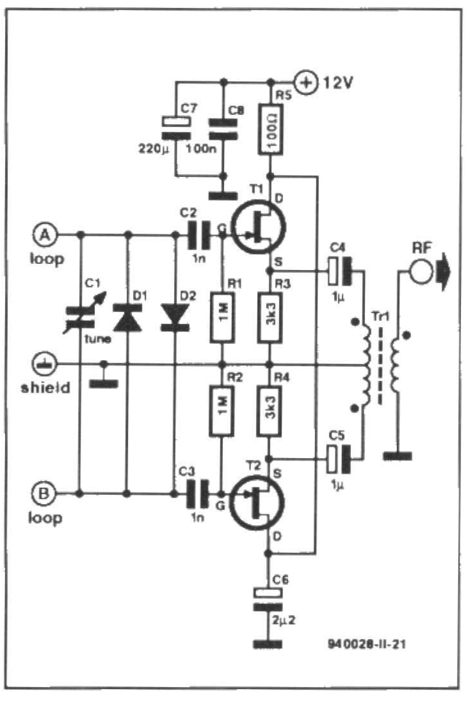


Fig. 23. Push-pull loop preamplifier.

teresting applications. The normal loop pattern is a figure-8 with very deep nulls being present at the broadside aspect to the antenna. The use in nulling interfering signals and in radio direction finding were described earlier. In this section we will describe two additional applications: the sports fan's loop and the monodirectional loop.

Sports fan's loop

This application uses a special form of loop antenna to boost the performance of AM BCB portable radios. It apparently originated when sports fans wanted to listen to ball games on distant AM stations that were normally out of the range of their portable receivers. The loop antenna is a square box loop, typically 60 to 150 cm on each side. A 100-cm square loop, with 8 turns of wire spaced to occupy 2.5 cm, produces an inductance of about 330 μH, which can be resonated to 550 kHz with 240 pF. No preamplifier is needed, although the resonating capacitor should be placed inside of a shielded metal box.

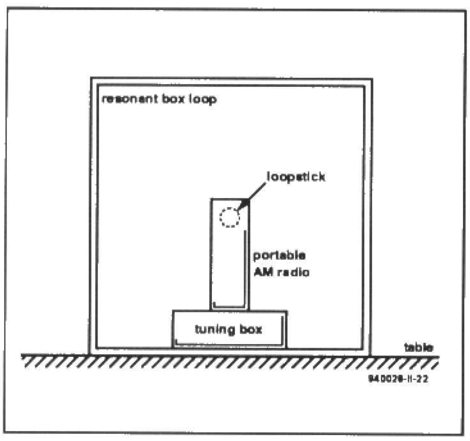


Fig. 24. Sports fans' loop.

The radio in the sports fan's loop is placed such that its internal loop stick receives in the same direction as the square loop (see **Fig. 24**). While a square loop has its nulls perpendicular (or broadside) to the loop, the loopstick has its nulls off the ends. Signals picked up by the larger loop are coupled into the loopstick antenna, providing a stronger signal for the radio to receive than is normally available with the loopstick alone.

At first, the version that I built (while researching my book *Joe Carr's Receiving Antenna Handbook*) did not work, so I wondered at the stories I had been told. However, the problem was soon found out: I had failed to know where the loopstick was inside the radio. I had assumed it was along the top of the radio, and ran from left-to-right relative to the front panel. However, it was actually mounted vertically along the right side of the radio near the tuning capacitor. Replacing the radio with one that had the loopstick antenna as shown in Fig. 24 solved the problem.

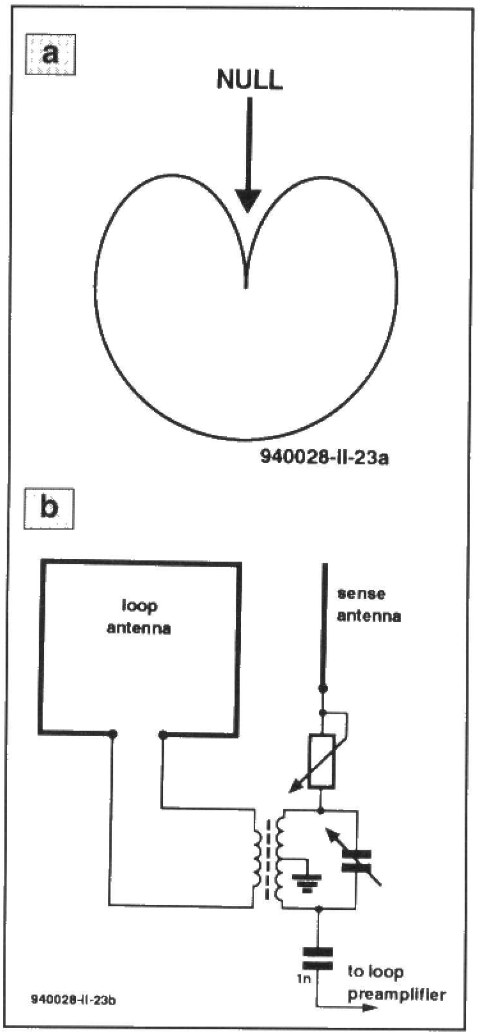


Fig. 25. (a) Monodirectional 'cardioid' pattern, (b) omni/monodirectional combination circuit.

Monodirectional reception

The normal loop pattern is bidirectional, figure-8, with deep nulls broadside to the loop. This pattern permits the antenna to null out, i.e. attenuate, any signal in the direction that the nulls point. Unfortunately, the bidirectionality causes two problems. In radio direction finding there is a directional ambiguity because RDF is typically done by pointing the null at the station until minimum signal level is achieved. The line perpendicular to the loop face contains the location of the station, but the station could be either in front of or behind the loop. In some cases, RDFers will take the directional measurement from three locations, and note where the three lines cross, which is the location of the station to a good precision. But in other cases, multiple location measurements are not feasible. In those applications, a monodirectional loop with but one null is needed.

Another problem caused by the loop is seen not in RDF, but when the loop user is at a fixed location that is on or near a line that runs between two stations. If you desire to listen to one of the stations, and the other is strong enough to cause interference, then nulling one with the loop also nulls the other. Assuming that the ratio of the signal levels is not such that nulling both places one below some comfortable threshold (a situation that I have never seen), one needs a monodirectional loop that has but one null.

There are two approaches to solving this problem. The classic approach is shown in **Fig. 25**. In this situation, the loop is paired with a small whip antenna used as a sense element, resulting in the cardioid pattern of **Fig. 25a**. The signal from the omnidirectional loop antenna is combined with the signal from the bidirectional loop antenna, in a network, as shown in **Fig. 25b**.

In use, the null is pointed at the offending station, while the maxima is pointed at the desired station.

The other approach is to use a spoiler loop in the manner of **Fig. 26** (Levintow n.d.). Here we have the undesirable situation of a pair of stations on the same or adjacent channels located such that the receiving site is on the line between the two stations. Two antennas are needed: the small ferrite loopstick and a 60 to 150 cm resonant box loop. The ferrite antenna can be a special antenna coupled to the receiver, e.g., when a receiver without an internal antenna is used, or it can be the normal loopstick inside of a portable receiver. The loopstick is placed such that it is broadside to the desired station. The box loop is placed 30 to 150 cm away from the loopstick, and in the direction of the offending station; the exact distance must be found experimentally for each situation. When the box loop is rotated through an angle of about 30 to 90 degrees with respect to the line between radio stations, a point will be found at which the offending station is nulled.

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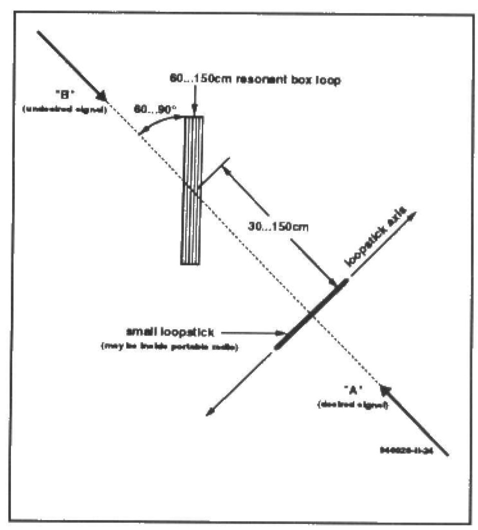

Fig. 26. Use of a secondary loop to null interference.

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COMPONENTS

Components for projects appearing in *Elektor Electronics* are usually available from appropriate advertisers in this magazine. If difficulties in the supply of components are envisaged, a source will normally be advised in the article. It should be noted that the source(s) given is (are) not exclusive — other suppliers may also be able to help.

BOOKS

For book availability, see advertisement on page 113.

SHELF BOX

Elektor Electronics shelf box ....£2.95...........$6.00

FRONT PANEL FOILS

PROJECT	No.	Price (£)	Price (US$)
4MB printer buffer card	920009-F	8.25	16.50
LC meter	920012-F	11.45	22.90
Guitar tuner	920033-F	8.80	17.60
NICAM decoder	920035-F	8.25	16.50
12VDC to 240VAC inverter	920038-F	16.15	32.30
Audio DAC	920063-F	10.00	20.00
Dig. audio/visual system	920022-F1	10.00	20.00
	920022-F2	19.40	38.80
	920022-F3	28.80	57.60
1.2 GHz multifunction frequency meter	920095-F	13.80	27.60
U2400B NiCd battery charger	920098-F	8.75	17.50
Workbench PSU	930033-F	17.00	34.00
Inexpensive phase meter	930046-F	17.25	34.50
Mini preamplifier	930106-F	Not available	
AF rms voltmeter	930108-F	17.25	34.50

EPROMS / PALS / MICROCONTROLLERS

PROJECT	No.	Price (£)	Price (US$)
EMON51 (8051 assembler course) (1 × 27256+disk 1661)	6061	20.00	40.00
EMON51 (8051 assembler course) (1 × 27256+disk 1681)	6091	20.00	40.00
FM tuner (1 × 27C256)	6101	15.30	30.60
Multi-purpose Z80 card: GAL set (2 × GAL 16V8)	6111	11.15	22.30
Multi-purpose Z80 card: BIOS (1 × EPROM 27128)	6121	15.30	30.60
1.2 GHz multifunction frequency meter(1 × 27C256)	6141	11.45	22.90
Digital audio/visual system (1 × 27C256)	6171	10.30	20.60
TV test pattern generator (1 × 27256)	6151	13.00	26.00
DiAV system. Package: 1 × 27512; 2 × GAL; 1 × floppy disk (MSDOS)	6181	30.50	61.00
PAL test pattern generator (1 × GAL 20V8-25)	6211	9.40	18.60
80C535 assembler course (1 x EMON52 EPROM and 1 x disk 1811)	6221	17.05	34.10
Watt-hour meter (1 × 27256)	6241	10.00	20.00
Four fold DAC (1 × GAL)	6251	10.75	21.50
Multipurpose display decoder (1 × 2764)	6261	11.50	23.00
Telephone-controlled switch (1 x 2764)	6271	14.50	29.00
Relative humidity meter (1 × 2764)	6301	14.50	29.00
535 card w. EPROM emulator (1 × PAL; 1 × GAL)	6311	26.00	52.00
Copybit eliminator (1 x GAL 16V8 and 1 x MACH110)	6321	42.25	84.50
RDS decoder (1 x 27C64)	6331	14.50	29.00
I²C tester (1 x GAL 6001)	6341	30.75	61.50
Mains signalling system (1 x 27C64)	6371	13.25	26.50
8751 programmer (1 × 8751)	7061	46.40	92.80
Microcontroller NiCd charger (1 × ST62E15)	7071	10.00	20.00
Maxi micro clock (clock)	7081	11.50	23.00
Maxi micro clock (darkroom timer)	7091	11.50	23.00
Maxi micro clock (cooking timer)	7101	11.50	23.00
Mini micro clock (clock)	7111	11.50	23.00
Mini micro clock (darkroom timer)	7121	11.50	23.00
Mini micro clock (cooking timer)	7131	11.50	23.00
VHF/UHF TV tuner (1 × 87C51)	7141	25.75	51.50
Microntroller-driven UART (1 × ST62T10)	7151	17.25	34.50
PIC programmer (1 x PIC17C42 and 1 x disk)	7161	52.75	105.50
MIDI swell pedal (1x27C64)	**946635**	**13.75**	**27.50**

DISKETTES

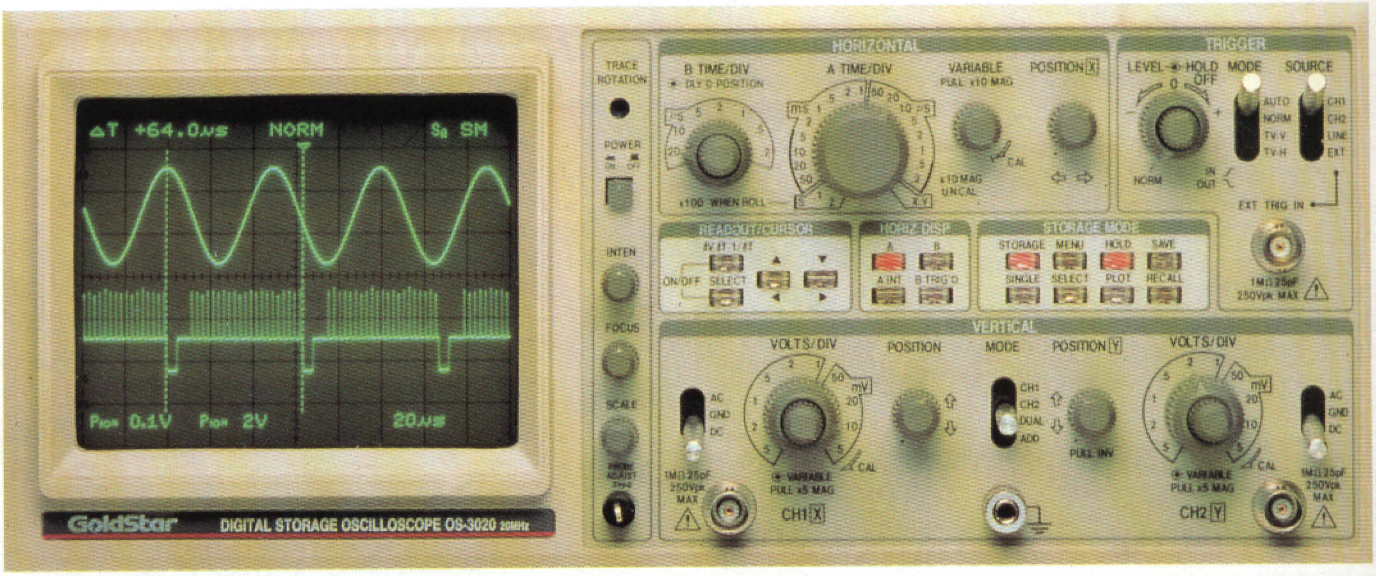
PROJECT	No.	Price (£)	Price (US$)
8051/8032 Assembler course (IBM version)	1661	7.65	15.30
8051/8032 Assembler course (Atari version) (3.5")	1681	7.65	15.30
AD232 converter	1691	7.65	15.30
GAL programmer (3 disks; upgrade: June 1993)	1701	11.15	22.30
Multi-purpose Z80 card	1711	7.65	15.30
Fuzzy Control One	1721	7.75	15.50
Pascal library for MMC	1751	9.70	19.40
Speech/sound memory	1771	7.65	15.30
PC-aided transistor tester	1781	7.50	15.00
IR receiver and DTMF decoder for 80C32 SBC	1791	9.00	18.00
80C535 assembler course	1811	8.80	17.60
I²C opto/relay card	1821	7.65	15.30
Video digitizer for PCs	1831	14.50	29.00
GAL programmer for Amiga	1841	11.00	22.00
I²C alphanumerical display	1851	8.50	17.00
Philips preamplifier	1861	8.50	17.00
GAL programmer (excl. Opal Jr. disks)	1881	10.75	21.50
Precision clock for PCs	1871	8.50	17.00
X2404-to-8751 interfacing	1891	8.50	17.00
RC5 software emulation	**1901**	**10.75**	**21.50**
Mains signalling system	1911	9.75	19.50
SIM — an 8051 simulator	1931	34.25	68.50
80C535 SBC extension	1941	9.75	19.50
80C535 SBC I²C display	946197-1	9.75	19.50
Small loop antennas	1951	10.75	21.50
RC5 transmitter w. 80C535	**946199-1**	**9.75**	**19.90**

PRINTED CIRCUIT BOARDS

Printed circuit boards whose number is followed by a ✛ sign are only available in combination with the associated software item, and can **not** be supplied separately. The indicated price includes the software.

PROJECT	No.	Price (£)	Price (US$)
FEBRUARY 1994			
80C535 single-board computer	924046	14.10	28.20
Copybit eliminator incl. MACH and GAL	930098✛	46.25	92.50
Mini preamplifier	930106	29.25	58.50
Bidirectional RS232-to-Centronics converter	930134	14.00	28.00
24 cm FM ATV transmitter	PCB available from author		
MARCH 1994			
PIC programmer incl. software set 7161	940048✛	66.00	132.00
100W AF amplifier			
- adaptor board	930039	8.25	16.50
- amplifier board	920135-1	19.40	38.80
- protection board	920135-2	7.95	15.90
Mini I/O card for Casio FX850/880P	930090	Not available	
AF signal tracer	940006	Not available	
Electronic fuse	936071	Not available	
APRIL 1994			
Mains signalling system - 1 (receiver)	940021-1	10.25	20.50
68HC11 processor board	930123	7.75	15.50
RS232 speedometer	930119	Not available	
Stroboscope light	940022	Not available	
Headphones amplifier	940016	18.75	37.50
Halogen light dimmer	940034	Not available	
RS232 interface for universal bus	924048	Not available	
MAY 1994			
Diffferential probe for oscilloscopes	940018	7.75	15.50
Mains signalling system - 2 (transmitter) (incl. disk 1911 and EPROM 6371)	940021-2✛	33.25	66.50
Morse decoder	contact Velleman distributor		
General-pupose sensor monitor	916107	Not available	
TV sound to light unit	936038	Not available	
JUNE 1994			
80C535 SBC extension	940025✛	19.75	39.50
I²C bus booster	940057	7.25	14.50
RS485 interface	940035	6.25	12.50
Fuel consumption monitor	940045	6.00	12.00
Intelligent EPROM eraser	940058-1	9.00	18.00
JULY/AUGUST 1994			
Gen. purpose IR volume control	930099	9.00	18.00
IR controlled switch	936066	9.75	19.50
MIDI swell pedal, incl. EPROM 946635	940019-C✛	26.75	53.50
Charge meter	940033	5.75	11.50
Water softener	944011-1	6.00	12.00
Joystick-to-mouse adaptor	944040-1	7.00	14.00
Discrete preamplifier	944063-1	8.50	17.00
Centronics I/O interface	944067-1	8.75	17.50
80C451 controller board	944069-1	15.00	30.00
Robust AF power amp	944075-1	9.75	19.50
PC over-temperature alarm	944076-1	4.25	8.50
1-4 MByte SIMM adaptor	944094-1	15.50	31.00
Optical doorbell	944080-1	6.25	12.50
PIC experimenting board	944105-1	17.75	35.50
RC5 transmitter w. 80C535 incl. disk 946199	944106-C✛	13.00	26.00
Car battery monitor	936042	Not available	
Resettable fuse for caravans	940061	Not available	

A list of all PCBs, software products and front panels available through the Readers Services is published in the March, June, September and December issues of *Elektor Electronics*.



Model OS-3020 illustrated above

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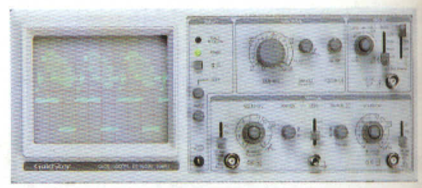
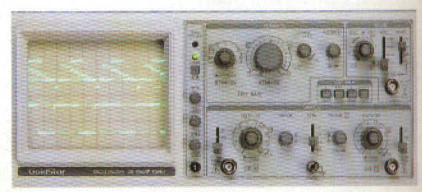


OS-9020A GW03D **£269.95**



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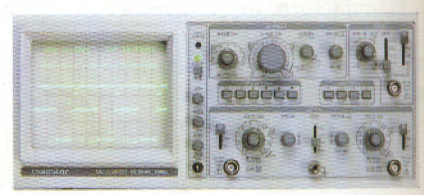

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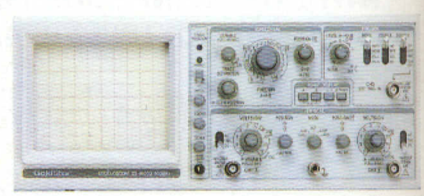
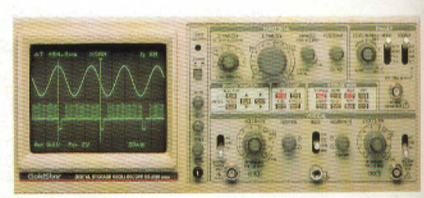


OS-9100D GW05F **£699.95**



OS-3020 GW07H **£749.95**

All models are supplied complete with probes, mains lead, spare fuses and detailed operating manual. Full details and specifications can be found in the **1994 Maplin Full Colour Catalogue**, available from **WHSMITH** and selected branches of **RSMcColl** in Scotland for £2.95 (£3.45 by post direct from Maplin). To order, Phone the Credit Card Hotline, 0702 554161, or send your Mail Order to P.O Box 3, Rayleigh, SS6 2BR, or visit your local Maplin store. Please note latest models all now cream in colour as Model OS-3020.

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